SENSORLESS SLIDING MODE VECTOR CONTROL OF THREE PHASE INDUCTION MOTOR DRIVES

A thesis submitted in partial fulfilments of the requirements for the award of the degree of

Master of Technology

in

ELECTRICAL ENGINEERING
(Industrial Electronics)

by

Prakash Behera
(Roll No – 213EE5340)

Department of Electrical Engineering
National Institute of Technology
Rourkela, Orissa

May, 2015

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Finally, I render my respect to my parents for giving me mental support and inspiration for carrying out my research work.

Date**:**

Place**:** Rourkela

Prakash Behera

M.Tech (Industrial Electronics)

National Institute of Technology Rourkela

CERTIFICATE

This is to certify that the work in this thesis entitled "**Sensorless Sliding Mode Vector Control of Three Phase Induction Motor Drives**" by **Prakash Behera**, Roll No.213EE5340, has been carried out under my supervision in partial fulfillment of the requirements for the degree of **Master of Technology** in Electrical Engineering with specialization "**Industrial Electronics**" during session 2014-2015 in the Department of Electrical Engineering, National Institute of Technology, Rourkela and this work has not been submitted elsewhere for a degree.

To the best of my knowledge and belief, this work has not been submitted to any other university or institution for the award of any degree or diploma.

Date **-**
Place **-** Rourkela

Prof. Kanungo Barada Mohanty
Department of Electrical Engineering
National Institute of Technology, Rourkela

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LIST OF SYMBOLS

$\omega_{\boldsymbol{r}}$	Rotor mechanical speed (rad/s)
ω_r^*	Reference or command speed (rad/s)
$\omega_{\boldsymbol{e}}$	Speed of the reference frame or synchronous speed (rad/s)
ω_{sl}	Electrical slip speed (rad/s)
Ψ_{dr}, Ψ_{qr}	d – axis and q – axis rotor flux linkages
$\Psi_{\alpha r}, \Psi_{\beta r}$	Component of rotor flux linkage vector in stationary (α-β) reference frame
$\Psi_{\alpha s}, \Psi_{\beta s}$	Component of stator flux linkage vector in stationary (α-β) reference frame
I_a^*, I_b^*, I_c^*	Reference current for abc phase of inverter
$T_{\boldsymbol{d}}$	Electromagnetic torque developed by motor
$T_{\boldsymbol{L}}$	Load torque
v_{ds}, v_{qs}	d and q – axis stator voltages
R_s, R_r	Resistance per phase of stator and rotor referred to stator
L_s, L_r	Stator and rotor per phase inductance referred to stator
L_m	Magnetizing inductance per phase referred to stator
D	Disturbance or noise in the sliding mode controller
E	Error vector in sliding mode controller
e	Error in a state (say, speed)
f	Supply frequency in Hz
G	Gain of sliding mode controller
J	Polar moment of inertia of the motor
K_T	Torque constant of the motor
S	Sliding surface
X	State vector
Y	Output vector
λ	Band width of the sliding mode controller
θ	Width of the boundary layer for the reduction of chattering
β	Viscous friction coefficient
ρ	Leakage coefficient
η	Positive constant used in sliding mode controller

ABSTRACT

In this thesis work, an indirect vector control of induction motor drive with a sliding mode controller is presented. The design consist of rotor speed estimation from measured stator terminal voltages and currents. The estimated speed is used as feedback in an indirect vector control of induction motor to achieving the speed control without the use of shaft mounted transducers. The specific contributions of the thesis are, first, a new sliding mode speed controller is developed by delicately introducing a switching variable and a design parameter. Then a fixed gain PI controller is also designed to compare the result. Combining the sliding mode controller and PI controller, the results are thoroughly analyzed to check the robustness of controller, reference speed tracking performance and regulator performance. Then, the robustness of the proposed sliding mode controller scheme is effectively demonstrated by considering standard inertia and high inertia load. Finally the speed is estimated by the rotor flux observer to realized sensorless control of induction motor drive.

CHAPTER-1

INTRODUCTION

1.1 Introduction

The induction motor has more preferences over separately excited dc motor in numerous perspectives, for example, performance, power to weight proportion, high speed capability, low starting expense, high dependability and robustness. Due to their basic and rigorous structure, high torque to weight proportion and can work in risky environment, induction motor can utilized in industrial drive applications. The dc motor can be control effectively because of they are decoupled as far as flux and torque. By utilizing vector control technique, the induction motor can be compelled to yield a quick torque response like that of a separately excited dc motor and it can be control linearly as dc motor by decoupling torque and flux commands.

In industrial drives, for elite speed control applications oblige four quadrant operation including field weakening, least torque ripple and fast speed recuperation under load torque and speed response. DC motor with thyristor converter and simple straight forward controller structure have been absolute choice for most modern and superior industrial applications. However, they had some certain issues like commutation requirement and maintenance. DC motor has some negative points as low torque to weight proportion and diminished unit capacity. Then again ac motors, particularly induction motors are suitable for mechanical drives because their simple and robust structure, high torque to weight proportion and high dependability. Yet, the control is very difficult on account of coupled impact of flux and torque component of current. On the off chance that it is conceivable if there should arise an occurrence of induction motor to control the amplitude and space angle (between rotating stator and rotor fields), intends to supply from a controlled source so that the flux and torque component of stator current can be control independently, so that motor dynamics can be compared to that dc motor with quick transient response. Instantly present of microcontroller, VLSI innovation and high switching frequency semiconductor gadgets like mosfets have directed to cost effective and simple control strategies.

1.1.1 Scalar control

The name scalar control demonstrates the extent variation of control variable only. For the control of speed of induction motor obliges a variable voltage variable frequency power sources. With the assistance of voltage source inverter (VSI), steady voltage/hertz (v/f) control etc can control the speed of induction motor on the grounds that it is basic, least expensive and hence forth it is a prominent strategy for speed control. In v/f control procedure keeping up the same terminal voltage to frequency proportion. So it give almost steady flux over extensive variety of speed variation. Since the flux is kept steady so the full load torque capacity are kept up consistent under enduring state condition with the exception of low speed (when an extra voltage is expected to compensate the stator winding voltage drop). In scalar control procedure, the steady state performance is great however transient performance is poor. In steady v/f control keeps the stator flux linkage consistent in steady state without keeping up decoupling between the flux and torque. So because of coupling effect transient response of the drive is poor. To stay away from open loop speed disturbance influence because of change in load torque and supply voltage, a close loop v/f speed control strategy with slip regulation is utilized for stable operation as a part of steady state [1].

Scalar control drives are generally utilized as a part of industry, on the grounds that is easy to actualized, yet there exists a coupling impact between both flux and torque (both are rely on up on voltage or current and frequency), which gives slow response and the system get to be instable. The significance of scalar control drives has lessened now a days in view of the considerable performance of the vector control drives.

1.1.2 Vector control

By splitting the stator current into two orthogonal components, one is towards flux linkage, which is magnetizing current or flux component of current, and other is perpendicular to the flux linkage, which is the torque component of current and the induction motor can be dealt with as a separately excited dc motor by changing both components autonomously. This ideas was concocted at the outset of 1970s. For the usage of vector control procedure includes data with respect to the magnitude and position of flux vector. Depending upon the method of flux acquisition, the vector control or field oriented control method can be divided into two types i.e. direct and indirect. In direct vector control the flux data can be find with the assistance of sensors or evaluated from machine terminal variables, for example, speed and stator current/voltage signals. Anyhow, the estimation of flux utilizing flux sensors obliged some assembling adjustment in the machines. Direct field orientation technique have its in conceived issue at low speed where the voltage drops because of resistances are more.

The indirect vector control is initially propose to takes out immediate estimation or computation of rotor flux from the machine terminal variables, however control its instantaneous flux position by including the rotor position signal with a command slip position signal. The direction of rotor position require a precise rotor speed data and the command slip position is computed from the induction motor model that again includes machine parameter which might change with temperature, frequency and also magnetic saturation.

1.2 Literature review

The vector control technique has been broadly utilized as a part of electric drive applications. By giving decoupled of torque and flux control commands, the vector control can like separately excited dc motor drive without changing the dynamic performance. In side with this scheme, a rotational transducer, for example tachogenerator or an speed encoder was frequently mounted to get the speed information. In this way, the speed information can be got and closed loop control is also completed. However the speed sensor may bring down the system dependability, expand the gadget investment. Therefore a sensorless speed control has been utilized modern industrial drive application.

By utilizing vector control strategy we can accomplish superior control of torque, speed or position from an industrial machine. This strategy can give same performance from an inverter driven induction machine as is reachable from a separately excited dc machine. Vector control gives decoupling control of rotor flux magnitude and torque producing current with quick torque response. The quick torque response is accomplished by evaluating, measuring or computing the magnitude and position of the rotor flux. Here indirect strategy for vector control is exhibited, the evaluation of rotor flux position is rely on upon rotor resistance. A parameter distinguishing algorithm, for example, the extended kalman filter can be utilized for the estimation of rotor resistance.

Induction motor drives are helpful for variable speed applications due to robustness of squirrel cage induction motor, their solved torque speed characteristics and advancement of high performance control algorithm. For the estimation of rotor speed, flux estimation is important. It is very difficult to measure flux using flux sensor. So there is a developing enthusiasm for evaluating rotor speed by utilizing just the stator voltage and current measurements.

A mid the most recent two decades, there have been made to replacement of conventional speed transducer in adjustable induction motor drives. By detecting the speed from electrical quantities like voltages and currents which are now accessible for the drive assessment and control. An early effort was made by Abbondanti and Brennen who designed in 1975 an analog slip calculator based on the processing of the motor input quantities, voltages, currents, and phases [5]. In 1979, it was followed by the work of Ishida et al. [11], who used rotor slot harmonic voltages in slip frequency control. They reported success for speeds over 300 r/min. Ishida and Iwata endorsed this technique in [12], [13], without any further progress in improving the speed range.

Latter, in 2003 S. Xepapas et.al [22] presented an observer which estimated the motor speed, rotor flux, the angular position of the rotor flux and the motor torque from measure terminal voltages and currents. They used nonlinear sliding mode control technique for both low and high speed control.

In 2005, M. Rashed et.al [14] proposed stable model reference adaptive system (MRAS) for speed estimation and designed a super twisting higher order sliding mode algorithm for speed control of induction motor drives which reduced the chattering problem.

Then in 2007, O. Barambones et.al [16] proposed Luenberger observer to estimating the rotor speed and used integral sliding mode controller for induction motor speed control based on field oriented control technique. C. Canvdas et.al [7] proposed sequential methodology for designing a robust adaptive sliding mode observer for an induction motor drive based on the singular perturbation theory.

1.3 Motivation

It is widely recognized that the induction motors which have been the traditional workhorse for fixed and variable speed drives, offer a very promising alternative for future applications like electric vehicles and robotics [7][15] and their utilities in industrial applications, e.g. web handling and overhead cranes, will also possibly increase over the next decade. So the control and estimation of induction motor drives organize a vast subject and the technology has further advanced in recent years. Hence, wide attentions of control researchers and engineers have been attracted to develop high performance control technique of ac drives.

On the other hand, the vector control technique have increased popularity in industrial applications in recent years, because they allow quick torque response like dc motor. There are two basic forms of rotor flux field orientations: direct field orientation in which relies on direct measurement or estimation of the rotor flux magnitude and angle and indirect field orientation which utilizes a slip relation of induction motor [8]. In direct vector control, air gap flux measurement has complex and lack of robustness. Thus the rotor speed can be measured, the rotor flux may be estimated based on the model of rotor flux equations. However, for the case of unknown rotor speed, estimation of the rotor flux is not an insignificant problem, especially at low speed range, there are problem due to variation of rotor parameters which are unavoidable due to the variation in the temperature and saturation levels of the machine [20][21]. The causes for the elimination of the speed sensors are that the speed sensors add cost, decrease reliability and noise problem. So the researcher try to estimate the speed without speed sensors.

We have been dedicated to developing simple implementation and reliable sensorless induction motor control and estimation technique by using the sliding mode control theory [25][26]. In the sensorless sliding mode control of induction motor, the rotor speed and rotor flux are estimated simultaneously.

1.4 Objective

The foremost objective of this research is to design and implementation of both sliding mode controller and PI controller to control the speed of induction motor without any speed sensors like tachogenerator and speed encoder.

The overall objective to be achieved in this study are:

- To design the equivalent d-q model of induction motor for its vector control analysis and closed loop operation of the drive system.
- Estimation of rotor speed of induction motor by using rotor flux observer to achieve sensorless control of the motor.
- Design and implementation of sliding mode controller and PI controller in MATLAB/Simulink software.
- Compare the simulation results using both controller in MATLAB/Simulink software to get their performance so as to consider better controller for our system applications.

1.5 Organization of the thesis

The thesis is organized as follows:

Chapter 1 shows the foundation for this thesis research, having presentation with a complete literature review in related area. Scalar control and vector control is described. Different kind of flux and speed estimation are additionally portrayed. The use of sliding mode controller to induction motor drives is introduced. Motivation an objective of my project works is also present.

Chapter 2 includes the mathematical modeling of induction motor. The differential equation of the motor are expressed in synchronously rotating (d – q) reference frame by taking stator current and rotor flux component as state variables. Basic field oriented control operation principles of induction motor like direct vector control and indirect vector control are briefly discussed. The rotor speed is estimated by using rotor flux observer to get sensorless control.

Chapter 3 describes the design of sliding mode controller for robust control of induction motor. The hypothesis of sliding mode control is quickly exhibited and control law is determined. The control gain and band width are dictated by different certainties like disturbances, rotor resistance variety and so on.

Chapter 4 presents the simulation of sliding mode controller and PI controller using MATLAB/Simulink software. Then compare the simulation result of both controller for checking robustness and parameter variation of motor. Then discussed trajectory tracking performance and regulator performance of sliding mode controller and sensorless speed estimation is also simulated.

At long last, in chapter 5, some conclusion on preferences, performance and constraint of the proposed controller are exhibited. This is likewise included proposals for future research work.

CHAPTER-2

MODELLING OF INDUCTION MOTOR

2.1 Introduction

Even if construction of induction motor is simple, its control of speed is considered to be more difficult as compare to DC motors, because it is nonlinear and highly interrelating multivariable state space model of motor. The quick changes and new improvement in microelectronics and power electronics variable frequency inverters with use of control hypothesis has made modern controllers for AC motor drives. The outline and utilization of such commute framework include fitting numerical modeling of the motor to upgrade the controller structure.

2.2 Modeling of induction motor

A suitable model for the three phase induction motor is required to simulate and study the complete induction motor drive system. The model of induction motor in rotating (d -q) reference frame is discussed in [2][17].

Assumptions taken for modeling of induction motor are listed below:

- Every stator winding is disseminated so as to generate a sinusoidal MMF along the airgap, i.e. space harmonics are ignored.
- The stator and rotor slots produces very negligible changes in inductances.
- Mutual inductances are equal.
- The voltages and currents harmonics are neglected.
- The magnetic circuit saturation is ignored.
- Hysteresis losses and eddy current losses and also skin effects are neglected.

The voltage equations of the three phase induction motor in synchronous rotating reference frame are [1][2]:

$$v_{ds} = R_s i_{ds} + \frac{d\Psi_{ds}}{dt} - \omega_e \Psi_{qs} \tag{2.1}$$

$$v_{qs} = R_s i_{qs} + \frac{d\Psi_{qs}}{dt} + \omega_e \Psi_{ds} \tag{2.2}$$

$$v_{dr} = R_r i_{dr} + \frac{d\Psi_{dr}}{dt} - (\omega_e - P\omega_r)\Psi_{qr} \tag{2.3}$$

$$v_{qr} = R_r i_{qr} + \frac{d\Psi_{qr}}{dt} + (\omega_e - P\omega_r)\Psi_{dr} \tag{2.4}$$

The develop torque T_d is

$$T_d = \frac{3}{2} P(\Psi_{ds} i_{qs} - \Psi_{qs} i_{ds}) \tag{2.5}$$

The torque balance equation is

$$J\frac{d\omega_r}{dt} = T_d - T_l - \beta\omega_r \tag{2.6}$$

Where voltages (v) and currents (i) refer to the synchronously rotating reference frame and the subscripts $ds, qs, dr\ and\ qr$ corresponds to d and q axis quantities for the stator and rotor respectively. Ψ is the flux linkages. ω_e is the speed of the reference frame and ω_r is the mechanical speed of the rotor in rad/sec. R_s is stator resistances per phase and R_r is rotor resistance per phase of the motor respectively. P is the number of pole pairs. J is the moment of inertia and β is the coefficient of viscous friction. T_d and T_l are the develop torque and the load torque respectively[2][23].

The above equations can be inscribed in matrix form as follows:

$$\begin{bmatrix} v_{ds} \\ v_{qs} \end{bmatrix} = \begin{bmatrix} R_s & 0 \\ 0 & R_s \end{bmatrix} \begin{bmatrix} i_{ds} \\ i_{qs} \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} \Psi_{ds} \\ \Psi_{qs} \end{bmatrix} + \begin{bmatrix} 0 & -\omega_e \\ \omega_e & 0 \end{bmatrix} \begin{bmatrix} \Psi_{ds} \\ \Psi_{qs} \end{bmatrix} \tag{2.7}$$

$$\begin{bmatrix} v_{dr} \\ v_{qr} \end{bmatrix} = \begin{bmatrix} R_r & 0 \\ 0 & R_r \end{bmatrix} \begin{bmatrix} i_{dr} \\ i_{qr} \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} \Psi_{dr} \\ \Psi_{qr} \end{bmatrix} + \begin{bmatrix} 0 & -(\omega_e - P\omega_r) \\ (\omega_e - P\omega_r) & 0 \end{bmatrix} \begin{bmatrix} \Psi_{ds} \\ \Psi_{qs} \end{bmatrix} \tag{2.8}$$

$$T_d = \frac{3}{2} P \begin{bmatrix} \Psi_{ds} & \Psi_{qs} \end{bmatrix} \begin{bmatrix} i_{qs} \\ -i_{ds} \end{bmatrix} \tag{2.9}$$

Squirrel cage induction motor is generally used so its rotor windings are short circuited.

$$\begin{bmatrix} v_{dr} \\ v_{qr} \end{bmatrix} = \begin{bmatrix} 0 \\ 0 \end{bmatrix} \tag{2.10}$$

Ignoring iron loss, then the flux linkage equations are in matrix form, as follows

$$\begin{bmatrix} \Psi_{ds} \\ \Psi_{qs} \end{bmatrix} = \begin{bmatrix} L_s & 0 \\ 0 & L_s \end{bmatrix} \begin{bmatrix} i_{ds} \\ i_{qs} \end{bmatrix} + \begin{bmatrix} L_m & 0 \\ 0 & L_m \end{bmatrix} \begin{bmatrix} i_{dr} \\ i_{qr} \end{bmatrix} \tag{2.11}$$

$$\begin{bmatrix} \Psi_{dr} \\ \Psi_{qr} \end{bmatrix} = \begin{bmatrix} L_m & 0 \\ 0 & L_m \end{bmatrix} \begin{bmatrix} i_{ds} \\ i_{qs} \end{bmatrix} + \begin{bmatrix} L_r & 0 \\ 0 & L_r \end{bmatrix} \begin{bmatrix} i_{dr} \\ i_{qr} \end{bmatrix} \tag{2.12}$$

Where L_s is self-inductances of stator and L_r is self-inductance of rotor respectively and also L_m is the mutual inductance.

From equation (2.12)

$$\begin{bmatrix} i_{dr} \\ i_{qr} \end{bmatrix} = \begin{bmatrix} \frac{1}{L_r} & 0 \\ 0 & \frac{1}{L_r} \end{bmatrix} \begin{bmatrix} \Psi_{dr} \\ \Psi_{qr} \end{bmatrix} - \begin{bmatrix} \frac{L_m}{L_r} & 0 \\ 0 & \frac{L_m}{L_r} \end{bmatrix} \begin{bmatrix} i_{ds} \\ i_{qs} \end{bmatrix} \tag{2.13}$$

Substituting equation (2.13) in (2.11)

$$\begin{bmatrix}\Psi_{ds}\\ \Psi_{qs}\end{bmatrix} = \begin{bmatrix}\left(L_s - \frac{L_m^2}{L_r}\right) & 0\\ 0 & \left(L_s - \frac{L_m^2}{L_r}\right)\end{bmatrix}\begin{bmatrix}i_{ds}\\ i_{qs}\end{bmatrix} + \begin{bmatrix}\frac{L_m}{L_r} & 0\\ 0 & \frac{L_m}{L_r}\end{bmatrix}\begin{bmatrix}\Psi_{dr}\\ \Psi_{qr}\end{bmatrix}$$

$$= \begin{bmatrix}\sigma L_s & 0\\ 0 & \sigma L_s\end{bmatrix}\begin{bmatrix}i_{ds}\\ i_{qs}\end{bmatrix} + \begin{bmatrix}\frac{L_m}{L_r} & 0\\ 0 & \frac{L_m}{L_r}\end{bmatrix}\begin{bmatrix}\Psi_{dr}\\ \Psi_{qr}\end{bmatrix} \qquad (2.14)$$

Where $\sigma = 1 - \frac{L_m^2}{L_s L_r}$ =Leakage coefficient.

Using equation (2.10) and substituting (2.13) in (2.8) then we get

$$\frac{d}{dt}\begin{bmatrix}\Psi_{dr}\\ \Psi_{qr}\end{bmatrix} = \begin{bmatrix}\frac{R_r L_m}{L_r} & 0\\ 0 & \frac{R_r L_m}{L_r}\end{bmatrix}\begin{bmatrix}i_{ds}\\ i_{qs}\end{bmatrix} + \begin{bmatrix}-\frac{R_r}{L_r} & 0\\ 0 & -\frac{R_r}{L_r}\end{bmatrix}\begin{bmatrix}\Psi_{dr}\\ \Psi_{qr}\end{bmatrix} + \begin{bmatrix}0 & -(\omega_e - P\omega_r)\\ (\omega_e - P\omega_r) & 0\end{bmatrix}\begin{bmatrix}\Psi_{dr}\\ \Psi_{qr}\end{bmatrix}$$

Or

$$\frac{d}{dt}\begin{bmatrix}\Psi_{dr}\\ \Psi_{qr}\end{bmatrix} = \begin{bmatrix}a_5 & 0\\ 0 & a_5\end{bmatrix}\begin{bmatrix}i_{ds}\\ i_{qs}\end{bmatrix} + \begin{bmatrix}-a_4 & \omega_{sl}\\ \omega_{sl} & -a_4\end{bmatrix}\begin{bmatrix}\Psi_{dr}\\ \Psi_{qr}\end{bmatrix} \qquad (2.15)$$

Where $a_5 = \frac{R_r L_m}{L_r}$, $a_4 = \frac{R_r}{L_r}$ and $\omega_{sl} = \omega_e - P\omega_r$ (2.15a)

From equation (2.14), (2.7) and (2.15), we get

$$\frac{d}{dt}\begin{bmatrix}i_{ds}\\ i_{qs}\end{bmatrix} = \begin{bmatrix}-a_1 & \omega_e\\ -\omega_e & -a_1\end{bmatrix}\begin{bmatrix}i_{ds}\\ i_{qs}\end{bmatrix} + \begin{bmatrix}a_2 & P a_3\omega_r\\ -P a_3\omega_r & a_2\end{bmatrix}\begin{bmatrix}\Psi_{dr}\\ \Psi_{qr}\end{bmatrix} + \begin{bmatrix}c & 0\\ 0 & c\end{bmatrix}\begin{bmatrix}v_{ds}\\ v_{qs}\end{bmatrix}$$

(2.16)

Where

$$a_1 = \frac{1}{\sigma L_s}\left(R_s + R_r\frac{L_m^2}{L_r^2}\right)$$

$$a_2 = \frac{1}{\sigma L_s}R_r\frac{L_m^2}{L_r^2}$$

$$a_3 = \frac{L_m}{\sigma L_s L_r} \qquad c = \frac{1}{\sigma L_s}$$

Relating equation (2.15) and (2.16), then we get the state space model of the induction motor as follows –

$$\frac{d}{dt}\begin{bmatrix}i_{ds}\\ i_{qs}\\ \Psi_{dr}\\ \Psi_{qr}\end{bmatrix} = \begin{bmatrix}-a_1 & \omega_e & a_2 & P a_3\omega_r\\ -\omega_e & -a_1 & -P a_3\omega_r & a_2\\ a_5 & 0 & -a_4 & \omega_{sl}\\ 0 & a_5 & -\omega_{sl} & -a_4\end{bmatrix}\begin{bmatrix}i_{ds}\\ i_{qs}\\ \Psi_{dr}\\ \Psi_{qr}\end{bmatrix} + \begin{bmatrix}c & 0\\ 0 & c\\ 0 & 0\\ 0 & 0\end{bmatrix}\begin{bmatrix}v_{ds}\\ v_{qs}\end{bmatrix} \qquad (2.17)$$

Using equation (2.14) and (2.9) and simplifying, we get

$$T_d = \frac{3}{2} P \frac{L_m}{L_r} [\Psi_{dr} \quad \Psi_{qr}] \begin{bmatrix} i_{qs} \\ -i_{ds} \end{bmatrix}$$

$$= \frac{3}{2} P \frac{L_m}{L_r} [\Psi_{dr} i_{qs} - \Psi_{qr} i_{ds}] \quad (2.18)$$

2.3 The Field Oriented control

The vector control technique is also known as field oriented control. It allows a squirrel cage induction motor to be driven with dynamic performance which is analogous to a dc motor. The field oriented control technique decouples the components of stator current; one providing the air gap flux and the other producing the torque.so that the control characteristic is linearized. The two component are d-axis i_{ds} corresponding to armature current, and q-axis i_{qs} corresponding to field current of a separately excited dc motor. The rotor flux linkage is allied along the d-axis of reference frame[24]. At this condition:

$$\Psi_{qr} = 0 \; and \; \Psi_{dr} = \Psi_r \quad (2.19)$$

The electromagnetic torque or developed torque of induction motor is given by [9][10]:

$$T_d = \frac{3}{2} P \frac{L_m}{L_r} \Psi_{dr} i_{qs} = K_t \Psi_{dr} i_{qs} \quad (2.20)$$

Where K_t and Piare the torque constant and number of pole pairs respectively.

With field orientation control technique the dynamic equation of stator current of d-q axis and rotor flux are given by:

$$\frac{d}{dt} i_{ds} = -a_1 i_{ds} + a_2 \Psi_{dr} + \omega_e i_{qs} + v_{ds}/(\sigma L_s) \quad (2.21)$$

$$\frac{d}{dt} i_{qs} = -a_1 i_{qs} - \omega_e i_{ds} - a_3 P \omega_r \Psi_{dr} + v_{qs}/(\sigma L_s) \quad (2.22)$$

$$\frac{d}{dt} \Psi_{dr} = -(\Psi_{dr}/\tau_r) + (L_m i_{ds}/\tau_r) \quad (2.23)$$

Where ω_e and ω_r are the synchronous speed (electrical) and the rotor speed (mechanical) in rad/s, v_{ds} *and* v_{qs} are d- axis and q- axis component of stator voltage,

$\sigma = 1 - \frac{L_m^2}{L_s L_r}$ is the leakage coefficient,

$\tau_r = {}^{L_r}/_{R_r}$ is rotor time constant,

$a_1 = \frac{R_s L_r + L_m^2/\tau_r}{\sigma L_s L_r}$, $a_2 = L_m/(\sigma L_s L_r \tau_r)$, $a_3 = L_m/(\sigma L_s L_r)$

Slip frequency for indirect vector control is

$$\omega_{sl} = \omega_e - P\omega_r = (L_m i_{qs})/(\tau_r \Psi_{dr}) \tag{2.24}$$

By decoupling of torque and flux, induction motor can be controlled linearly like separately excited dc motor. But, due to the rotor time constant τ_r in equation (2.24), the indirect field oriented control is highly parameter sensitive.

Vector control technique can be implemented with reference to any of the flux vectors: like stator flux, air gap flux and rotor flux. But out of the three the rotor flux orientation gives a natural decoupling between flux and torque, fast torque response as well as better stability. Hence in this work rotor flux orientation is considered.

Depending up on how the unit vector is generated, the field oriented control is divided in to two types and they are

(a) Direct or feedback vector control

(b) Indirect or feedforward vector control

2.3.1 Direct or feedback vector control

In direct vector control the unit vector is generated from rotor flux vector which is obtained from direct measurement of airgap flux component or estimated from direct measurement of terminal quantities like stator current, stator voltage or speed.

Direct vector control can be achieved by following method:

(1) Sensing airgap flux component using flux sensors.

(2) Sensing stator voltage and current and estimating rotor flux using voltage model.

(3) Sensing stator current and rotor speed and estimating rotor flux using current model.

The direct method of vector control is difficult to operate successfully at very low frequency (including zero speed) because of following problems:

(1) At low frequency, voltage signals $v_{\alpha s}$ and $v_{\beta s}$ are very low and ideal integration becomes difficult because dc offset tends to build up at integrator output.

(2) The parameter variation effect of resistance and inductance tends to reduce accuracy of estimated signals. Particularly temperature variation of resistance become more dominant.

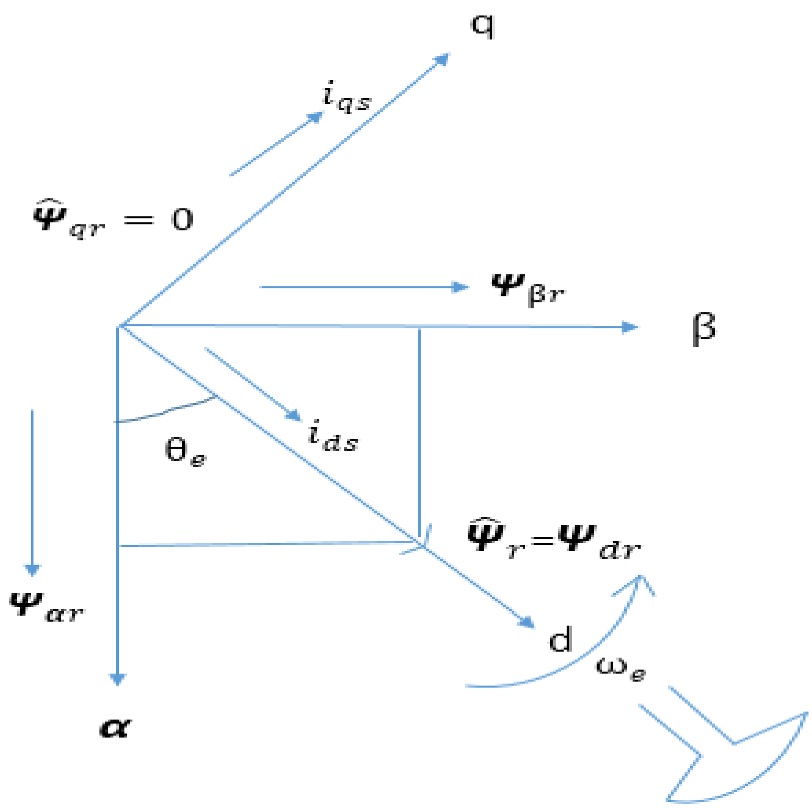


Fig. 2.1 Phasor diagram explaining direct vector control

2.3.2 Indirect or feedforward vector control

The indirect vector control method is essentially the same as direct vector control, except the unit vector signals are generated in feedforward manner. The $\alpha - \beta$ axes are fixed on the stator, but $d_r - q_r$ axes, which are fixed on rotor, are moving at the speed ω_r. Synchronously rotating axes $d - q$ are rotating ahead of the $d_r - q_r$ axes by the positive slip angle θ_{sl} corresponding to slip frequency ω_{sl}. Since the rotor pole is directed on the d axis and $\omega_e = \omega_r + \omega_{sl}$, we can write [1]

$$\theta_e = \int \omega_e dt = \int (\omega_r + \omega_{sl}) dt = \theta_r + \theta_{sl}$$

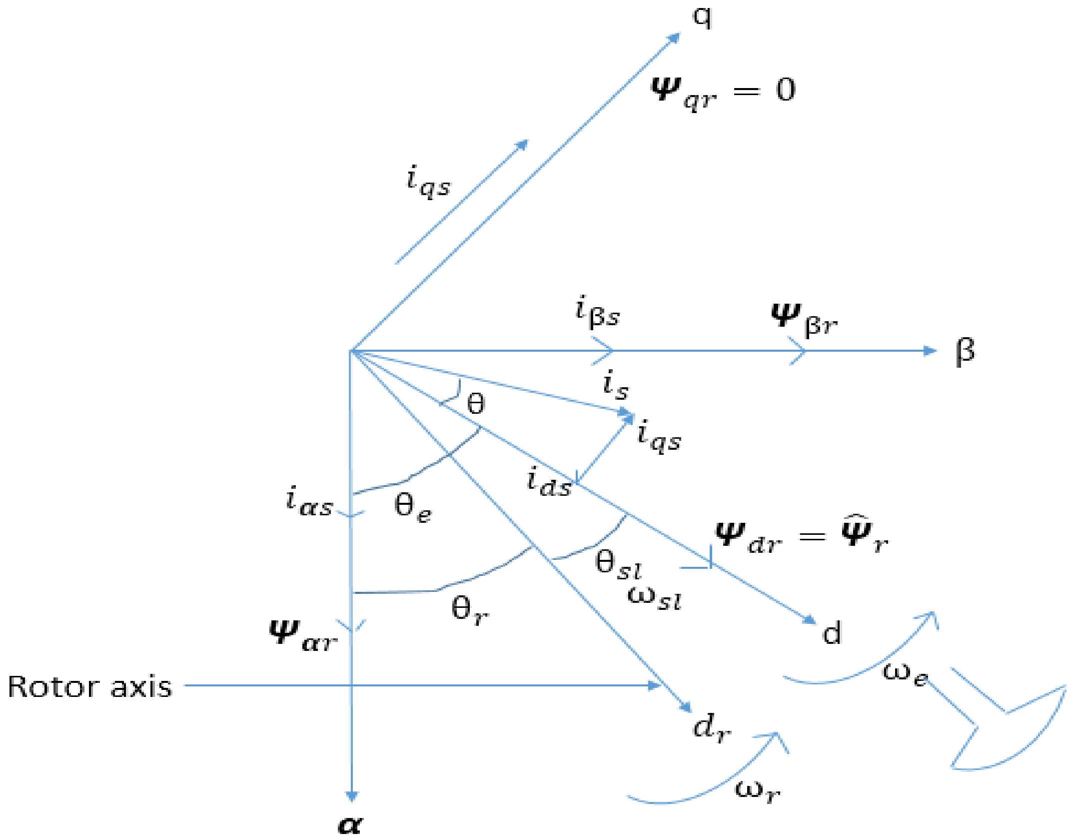


Fig. 2.2 Phasor diagram explaining indirect vector control

2.4 Estimation of Speed

Sensorless vector control of induction motor drive means vector control without speed sensor. An incremental shaft mounted speed encoder is required for closed loop speed control. A speed encoder is undesirable in a drive because it adds cost and reliability problems, also the need for a shaft extension and mounting arrangement [1][2].

Induction motor speed estimation has following technique:

(a) Slip calculation

(b) Direct synthesis from rotor flux observer

(c) Model reference adaptive system (MRAS)

(d) Speed adaptive flux observer (Luenberger observer)

(e) Extended Kalman filter

(f) Slot harmonics

(g) Injection of auxiliary signal on salient rotor

2.4.1 Speed estimation using rotor flux observer

The greater part of the speed estimation strategies are in light of streamlined motor models. Keeping in mind the end goal to show signs of improvement speed estimation, it is obliged to have dynamic representation in view of the stationary reference frame. Since motor voltages and currents are measured in a stationary edge, so it is suitable to express these mathematical equation in stationary reference outline [9][10][22].

From the stator voltage equations in stationary reference frame:

$$v_{\alpha s} = R_s i_{\alpha s} + \frac{d}{dt}\Psi_{\alpha s} \tag{2.25}$$

We know

$$\begin{aligned} \Psi_{\alpha s} &= L_s i_{\alpha s} + L_m i_{\alpha r} \\ &= L_{ls} i_{\alpha s} + L_m(i_{\alpha s} + i_{\alpha r}) \\ &= L_{ls} i_{\alpha s} + \Psi_{\alpha m} \end{aligned} \tag{2.26}$$

Put equation (2.26) in equation (2.25) we get

$$v_{\alpha s} = R_s i_{\alpha s} + L_{ls}\frac{d}{dt} i_{\alpha s} + \dot{\Psi}_{\alpha m} \tag{2.27}$$

$$\Psi_{\alpha m} = L_m(i_{\alpha s} + i_{\alpha r})$$

$$\Longrightarrow \frac{L_r}{L_m}\Psi_{\alpha m} = L_{lr} i_{\alpha s} + \Psi_{\alpha r}$$

$$\Longrightarrow \Psi_{\alpha m} = \frac{L_m}{L_r} L_{lr} i_{\alpha s} + \frac{L_m}{L_r}\Psi_{\alpha r} \tag{2.28}$$

Substitute equation (2.28) in equation (2.27) and multiply with $\frac{L_r}{L_m}$ we get [3][10]

$$\frac{d}{dt}\Psi_{\alpha r} = \frac{L_r}{L_m} v_{\alpha s} - \frac{L_r}{L_m}(R_s + \sigma L_s S) i_{\alpha s} \tag{2.29}$$

Similarly

$$\frac{d}{dt}\Psi_{\beta r} = \frac{L_r}{L_m} v_{\beta s} - \frac{L_r}{L_m}(R_s + \sigma L_s S) i_{\beta s} \tag{2.30}$$

Where Ψ_r is the rotor flux linkage. v_s and.i_s are stator voltage and current respectively. α and β denote the corresponding components of rotor flux linkage or stator voltage or current.

Eliminating current derivatives from equation (2.29) and (2.30), the rotor flux dynamics equations in the stationary frame are obtained as follows:

$$\dot{\Psi}_{\alpha r} = (L_m/\tau_r) i_{\alpha s} - P\omega_r \Psi_{\beta r} - \Psi_{\alpha r}/\tau_r \tag{2.31}$$

$$\dot{\Psi}_{\beta r} = (L_m/\tau_r) i_{\beta s} - P\omega_r \Psi_{\alpha r} - \Psi_{\beta r}/\tau_r \tag{2.32}$$

The electrical angle θ_e of the rotor flux vector ($\overline{\Psi}_r$) with the α-axis of the stationary reference frame is given by[18]:

$$\theta_e = tan^{-1}(\Psi_{\beta r}/\Psi_{\alpha r}) \tag{2.33}$$

The time derivative of flux angle is the instantaneous electrical synchronous speed.

$$\omega_e = \dot{\theta}_e = (\Psi_{\alpha r}\dot{\Psi}_{\beta r} - \Psi_{\beta r}\dot{\Psi}_{\alpha r})/\Psi_r^2 \tag{2.34}$$

Substituting equation (2.29) and (2.30) in equation (2.34), then we get

$$\omega_e = P\omega_r + (\Psi_{\alpha r}i_{\beta s} - \Psi_{\beta r}i_{\alpha s})L_m/(\tau_r\Psi_r^2) \tag{2.35}$$

Solving equations (2.34) and (2.35), then we find estimated speed as follows [1][18][19],

$$\widehat{\omega}_r = \left[\Psi_{\alpha r}\dot{\Psi}_{\beta r} - \Psi_{\beta r}\dot{\Psi}_{\alpha r} - \left(\Psi_{\alpha r}i_{\beta s} - \Psi_{\beta r}i_{\alpha s}\right)(L_m/\tau_r)\right]/(P\Psi_r^2)$$

2.5 Conclusion

The mathematical modeling of induction motor is discussed. The differential equation of the motor are expressed in synchronously rotating (d – q) reference frame by taking stator current and rotor flux component as state variables. Basic field oriented control operation principles of induction motor like direct vector control and indirect vector control are briefly discussed. The rotor speed is estimated by using rotor flux observer to get sensorless control. By using this method, the speed estimation is highly machine parameter sensitive and have a tendency to give poor accuracy of rotor speed estimation

CHAPTER-3

DESIGN OF A SLIDINGMODE CONTROLLER

3.1 Introduction

A variable sliding mode control structure is mostly an adaptive nonlinear control that gives robust performance to parameter variation as well as load torque disturbance. It can be applied to a linear or nonlinear plant. By using sliding mode control, the drive response is forced to tract or slide along a predefined trajectory or reference model by a switching control algorithm, despite of the system's parameter variation along with load disturbance. The control DSP detects the deviation of the actual trajectory from the reference trajectory and accordingly the switching strategy is also changing to restore the tracking.

The advantages of sliding mode control is:

- System behaves like reduced order system.
- Robust to parameter changes, uncertainties and disturbances.
- Due to discontinuous control law, convergence time is finite.

The disadvantages of sliding mode control is:

- High frequency chattering
- Sometimes actuator cannot response at that fast due to inertia.
- Structure cost is high due to discontinuous input.

The design of sliding mode control comprises two steps:

- Choice of stable hyper plane(s) in the state/error space on which motion should be controlled, called the switching function.
- The selected sliding surface becomes attractive by using the appropriate design of control law.

3.2 Sliding mode controller

By utilizing sliding mode controller, the system is controlled so that the deviation in the system dependably ventures towards a sliding surface. The sliding surface has two variables such as one is the tracking error (e) of the state and another is the rate of change ($\dot{e}$). The control input to the system can be choose by utilizing the separation of the error trajectory from the sliding surface and its rate of union. The indication of the control input need to change at the crossing point of the following error trajectory with the sliding surface. Thusly the error trajectory is constantly implemented to move towards the sliding surface.

The equations (2.20 – 2.24) of chapter - 2 are simplified by supposing, the rotor flux Ψ_{dr} to be constant. From (2.23) the steady state value of the rotor flux can be got as [13][23][19]

$$\Psi_{dr}^{*} = \frac{a_5}{a_4} i_{ds}^{*} \tag{3.1}$$

From (2.15a), we get

$$\frac{a_5}{a_4} = L_m \tag{3.1a}$$

Substitute equation (3.1a) in (3.1) we get

$$\Psi_{dr}^{*} = L_m i_{ds}^{*} \tag{3.1b}$$

The speed dynamic equation is given by

$$\frac{d}{dt}\omega_r = \frac{1}{J}(T_e - T_l - \beta\omega_r) \tag{3.2}$$

Using equation (2.20) in (3.2) we get

$$\dot{\omega}_r = \frac{1}{J}(K_t \Psi_{dr}^{*} i_{qs} - \beta\omega_r - T_l)$$

Suppose the lode torque, T_l is disturbance to a system, the speed dynamic equation is written as:

$$\dot{\omega}_r = -\frac{\beta}{J}\omega_r + b i_{qs} + noise \tag{3.3}$$

or

$$\dot{\omega}_r = f_1 + noise \tag{3.3a}$$

Where $f_1 = -\frac{\beta}{J}\omega_r + b i_{qs}$ (3.4)

and $b = \frac{K_t}{J}\Psi_{dr}^{*}$ (3.4a)

Speed is used as the output variable in field oriented controlled induction motor drive. To track the speed perfectly in the second order speed control system, the conditions to be fulfilled are [22][23]:

$$\dot{\omega}_r \, I_{\omega_r = \omega_r^{*}} = 0 \qquad and \qquad \ddot{\omega}_r \, I_{\omega_r = \omega_r^{*}} = 0 \tag{3.5}$$

From equation (3.3)

$$\ddot{\omega}_r = -\frac{\beta}{J}\dot{\omega}_r + b\frac{d}{dt} i_{qs} + noise \tag{3.5a}$$

Substituting $\omega_e = P\omega_r + a_5 \frac{i_{qs}}{\Psi_{dr}}$ in equation (2.22), the following equation is obtained.

$$\frac{d}{dt} i_{qs} = -\left(P\omega_r + a_4 \frac{i_{qs}}{i_{ds}}\right) i_{ds} - a_1 i_{qs} - P a_3 \omega_r L_m i_{ds} + c v_{qs} \tag{3.6}$$

It can be shown using equation (2.24) and (3.1), that

$$\omega_{sl} = a_5 \frac{i_{qs}}{\Psi_{dr}^*} = \frac{R_r L_m}{L_r} \frac{i_{qs}}{L_m i_{ds}^*} \approx a_4 \frac{i_{qs}}{i_{ds}}$$

Simplifying equation (3.6), we get

$$\frac{d}{dt} i_{qs} = -(a_1 + a_4) i_{qs} - P\omega_r (1 + a_3 L_m) i_{ds} + c v_{qs}$$

or

$$\frac{d}{dt} i_{qs} = f_2 + c v_{qs} \tag{3.6a}$$

Where

$$f_2 = -(a_1 + a_4) i_{qs} - P\omega_r (1 + a_3 L_m) i_{ds} \tag{3.7}$$

Substituting equation (3.2) and (3.6a) in equation (3.5a), we get

$$\ddot{\omega}_r = -\frac{\beta}{J} f_1 + b f_2 + b c v_{qs} + d$$

Or

$$\ddot{\omega}_r = G + u + d \tag{3.8}$$

Where d = total disturbance

u = bcv_{qs}=control input (3.8a)

$$G = -\frac{\beta}{J} f_1 + b f_2 \tag{3.9}$$

G is a function that can be estimated from measured values of current and speed.

u is directly proportionate to v_{qs} which decides the controlling signal.

Let $G = \hat{G} + \Delta G$ (3.10)

Where $\hat{G}$ is an approximate of G and ΔG is the estimation error due to modeling imperfection.

The control problem is to obtain the system states,

$$X = [\omega_r \quad \dot{\omega}_r]^T$$

To track a specific time varying state,

$$X^* = [\omega_r^* \quad \dot{\omega}_r^*]^T$$

In presence of modeling imperfection and disturbances.

Let $e = \omega_r - \omega_r^*$ (3.11)

and $\dot{e} = \dot{\omega}_r - \dot{\omega}_r^*$ (3.12)

be the tracking error in the speed and its rate of change respectively.

Let $E = [e \quad \dot{e}]^T$ be the tracking error vector.

A time varying surface, S(t) is defined in the state space by the scalar equation,

$$S(x,t) = 0$$

Where $$S(x,t) = \left(\frac{d}{dt} + \lambda\right) e = \dot{e} + \lambda e \tag{3.13}$$

Where λ is a positive constant, which determines the band width of the system.

Starting from the initial condition, E(0)=0, the tracking task, $X \to X^*$, which means X has to follow $X^*$with a predefined trajectory , is considered as solved, if the state vector, E remains in the sliding surface, S(t) for all t≥0 and scalar quantity S is had at zero. A sufficient condition for this behavior is to choose the control law, u of equation (3.8) and (3.8a) so that [13] [18][19]

$$\frac{1}{2}\frac{d}{dt}(S^2) \leq -\eta|S| \tag{3.14}$$

Where η is a positive constant. The value of η decides the degree to which the system state is attracted to the switching line. Equation (3.14) states that the squared "distance" to the sliding surface, as measured by S^2, decreases along all system trajectories.

For a second order system, the sliding surface is a line and diverse control structures are connected by position of the following error vector concerning the line. The control is intended to drive the state into the switching line, and once in the line the system state is considered to stay hanging in the balance. The state trajectory is characterized by the arithmetical comparison of the line.

In sliding mode, conduct of the system is invariant notwithstanding modeling blemishes, parameter variety and aggravations. However the sliding mode causes intense changes of the control variable, which is a noteworthy downside for the system.

3.2.1 Derivation of control input

The control problem is to get the motor speed ω_r to track a specific time varying command ω_r^* in presence of model inaccuracy, load torque disturbances and measurement noise. In sliding mode control, the system is controlled in such way that the tracking error, $e = \omega_r - \omega_r^*$ and its rate of change $\dot{e}$ always move towards a sliding surface. The sliding surface is defined in the state space by the scalar equation [18][16]:

$$S(e, \dot{e}, t) = 0$$

Where sliding variable, S is:

$$S = \dot{e} + \lambda e \tag{3.15}$$

Where, λ is positive constant that depends on the bandwidth of the system.

The issue of following is proportional to staying on the sliding surface for unsurpassed, and the sliding variable, S is kept at zero. For a second order system the sliding surface is a line. The separation of the error trajectory from the sliding surface and its rate of joining are utilized to choose the control input. The indication of control input must change at the crossing point of following error trajectory with the sliding surface. In this way the error trajectory is forced to move always towards the sliding surface. The condition of sliding mode is[23][9][10]:

$$\frac{1}{2}\frac{d}{dt}S^2 = S\dot{S} \leq -\eta|S| \qquad (3.16)$$

Where η is positive constant.

The equation (3.16) is stricter than the general sliding condition: $S\dot{S} \leq 0$

Equation (3.16) is equivalent to

$$\dot{S}sgn(S) \leq -\eta \qquad (3.17)$$

To design a slidin mode speed controller for the field oriented induction motor drive system, the steps are as follows. Substituting equation (3.15) in (3.17), we get:

$$(\ddot{\omega}_r + \lambda\dot{e} - \ddot{\omega}_r^*)\ sgn(S) \leq -\eta \qquad (3.18)$$

The speed dynamics is given by:

$$J\dot{\omega}_r + B\omega_r + T_L = T_e = K_T\Psi_{dr}i_{qs} \qquad (3.19)$$

Equivalently, we can express equation (3.19) as:

$$\dot{\omega}_r = g_1 - ({}^{T_L}/_J) \qquad (3.20)$$

Where,

$$g_1 = (-B\omega_r + K_T\Psi_{dr}i_{qs})/J$$

Differentiating equation (3.20) with respect to time and simplifying, we get:

$$\ddot{\omega}_r = G + d + bi_{qs} \qquad (3.21)$$

Where,

$$G = (-Bg_1 + K_T\Psi_{dr}^*g_2)/J \qquad (3.22)$$

$$g_2 = -(a_1 + \tau_r^{-1})i_{qs} - P\omega_r(1 + a_3L_m)i_{ds}$$

$$b = K_T\Psi_{dr}^*/(\sigma L_sJ)$$

G is a function which can be evaluated from measured estimations of currents and speed and d is unsettling influence because of the load torque, and error in estimation of G, which may happen because of estimation errors. Third term is the control term, as the q-axis stator current command is in charge of evolving torque. In the most essential sliding mode controller, no evaluation or estimation is finished. It doesn't look into G and d. It is defined as:

$$i_{qs}^{*} = -K.sgn(S) \tag{3.23}$$

Where K is a positive constant, which is the gain of the sliding mode controller.

To improve the performance, Slotine proposed the sliding mode controller, which is obtained as follows.

Substitute equation (3.21) in (3.18) and simplifying:

$$(G + d + \lambda\dot{e} - \ddot{\omega}_r^{*})sgn(S) + bi_{qs}sgn(S) \leq -\eta \tag{3.24}$$

From equation (3.23) and (3.24) dual component sliding mode controller without boundary layer is obtained as[9]:

$$i_{qs}^{*} = \left\{\frac{-\hat{G}-\lambda\dot{e}+\ddot{\omega}_r^{*}}{\hat{b}}\right\} - Ksgn(s) \tag{3.25}$$

The first term of equation (3.25), is a compensation term and the second term is the controller. The compensation term is persistent and reflects information of the system elements. The controller term is intermittent and guarantees the sliding to happen.

3.2.2 Reduction of chattering

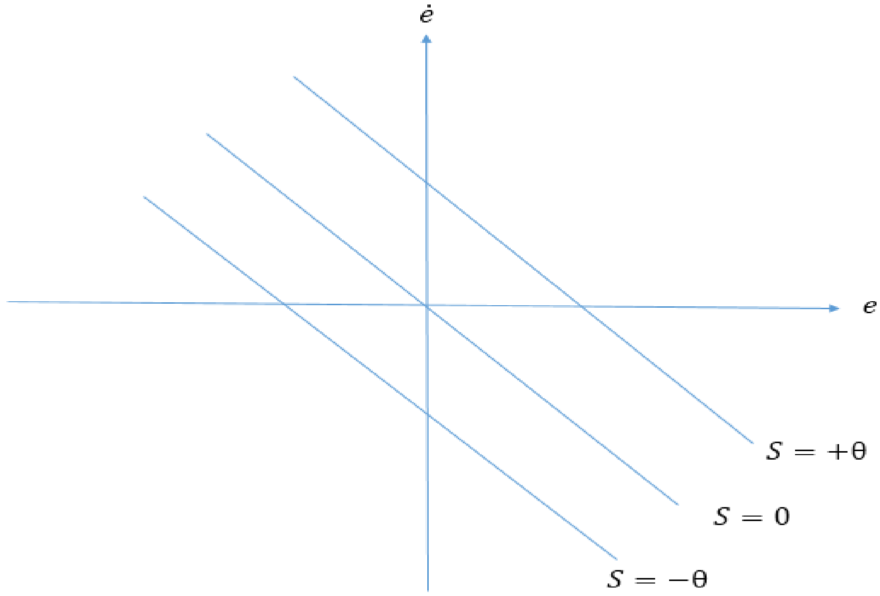
Despite the fact that dual component sliding mode controller given by comparison (3.25) has preferred performance over single component sliding mode controller given by mathematical statement (3.23), yet some sum prattling is available. To lessen gabbing a limit layer of width θ is presented on both sides of the sliding line. This adds up to lessening of the control gain inside the limit layer and results in a smooth control sign. Then the control law of equation (3.23) modifies to:

$$i_{qs}^{*} = -Ksat(S/\theta) \tag{3.26}$$

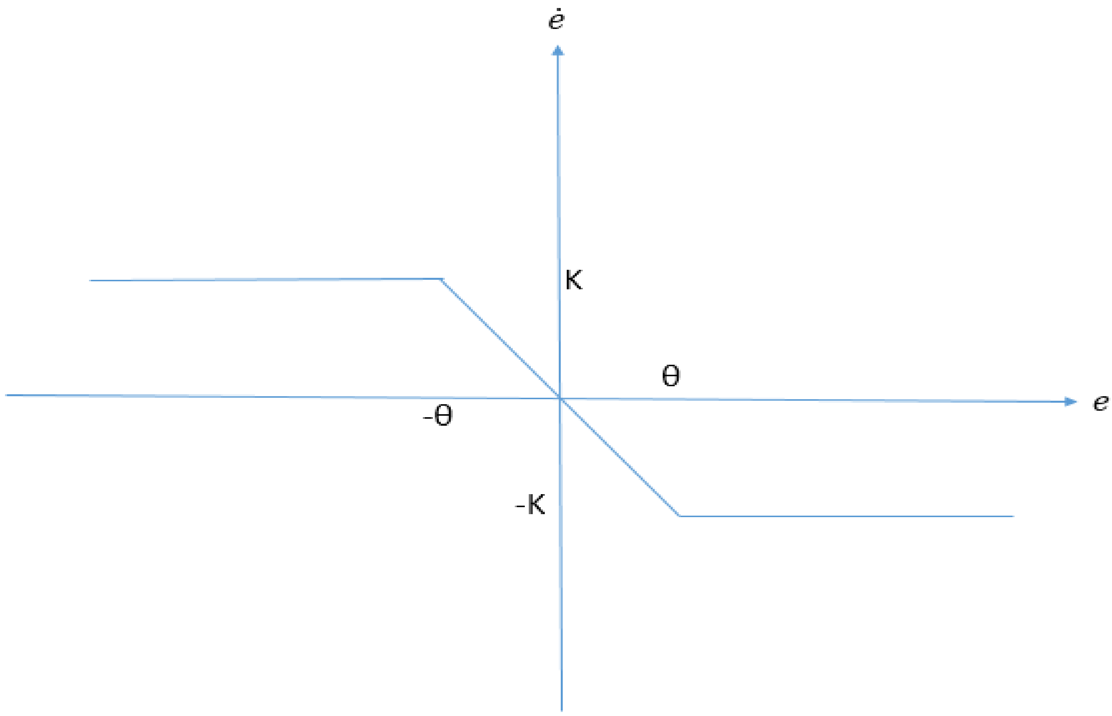
Where $$sat(S/\theta) = \begin{cases} S/\theta & if\ |S| \leq \theta \\ sgn(S) & if|S| > \theta \end{cases}$$

To reduce this chattering, sliding mode controller with boundary layer as given by following equation [9][10]:

$$i_{qs}^{*} = \left\{\frac{-\hat{G}-\lambda\dot{e}+\ddot{\omega}_r^{*}}{\hat{b}}\right\} - Ksat(S/\theta) \tag{3.27}$$



(a)



(b)

Fig.3.1 sliding mode principle with boundary layer

3.3 Conclusion

The outline and usage of sliding mode controller is depicted for vigorous control of induction motor. The hypothesis of sliding mode control is quickly exhibited and control law is inferred. The controller gain and band width are dictated by different realities like load changes, rotor resistance variety and so on. The control law is changed to lessen the chattering effect.

CHAPTER-4

SIMULATION RESULTS AND DISCUSSIONS

4.1 Introduction

The 3-phase induction motor drive system, whose rating and parameters are given in appendix section, is recreated with both sliding mode controller and PI controller. Both the controllers are tried for speed tracking and load torque variety conditions. The sliding mode controller is tested for robustness with two different moment of inertia values:

Standard inertia, J_{min}=0.0088Kg. m^2 and

High inertia, J_{max}=0.0167 Kg. m^2

The relative investigation of the consequences of SMC and PI controller are demonstrated as follows:

4.2 Trajectory tracking performance

If there should be an occurrence of trajectory tracking utilizing sliding mode control, if the error trajectory moves far from the beginning of the phase plane, it return by being driven on to the sliding line and then sliding along it. Inside of the boundary layer the error trajectory is not driven on to the switching line. Error trajectory having entered the boundary layer are obliged to move inside of it. The specific type of the intermittent reference speed trajectory utilized for tests is demonstrated as a part of Fig.4.1. The command speed is increased linearly from 0 at t=0.2s to 90 rad/s at t=0.5s. It is kept constant at 90 rad/s till t=1.5s, and decreased linearly to -90 rad/s at t=2.1s. Then command speed is kept constant at -90 rad/s till t=3.1s and increased linearly to 0 at t=3.4s. Load torque of 3 Nm (rated torque) is applied from t=0.75s to t=1.25s and -3 Nm is applied from t=2.35s to t=2.85s. The same trajectory is used for both sliding mode controller and P-I controller and results are compared.

The trajectory following result of sliding mode controller is presented for two different moment of inertia, i.e. with standard inertia and high inertia load, the reference speed and the actual rotor speed trajectory is shown in Fig.4.3. and speed error response is shown in Fig.4.4 for both value of inertia. Fig.4.5 shows the control output which is the reference q-axis stator current with stand inertia load. This result demonstrates the robustness of the drive system with sliding mode controller. With high inertia load, sliding mode controller gives less error as compare to standard inertia load. Fig.4.6 shows the variation of switching variable for sliding mode implementation. Speed error and control output for P-I controller with standard inertia load and high inertia load are shown in Fig.4.8 and Fig.4.9.Thus sliding mode controller gives superior performance compared to P-I controller for tracking fast changing trajectory. It also gives robust control against parameter variations.

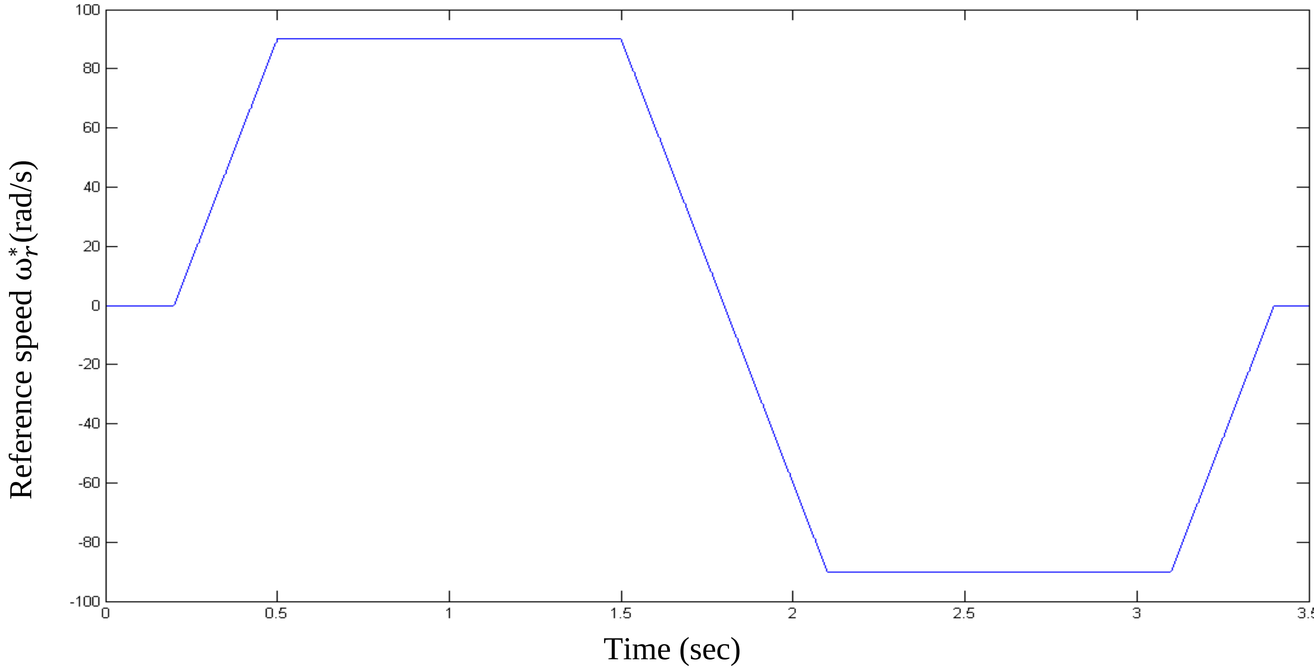


Fig. 4.1 Reference speed trajectory

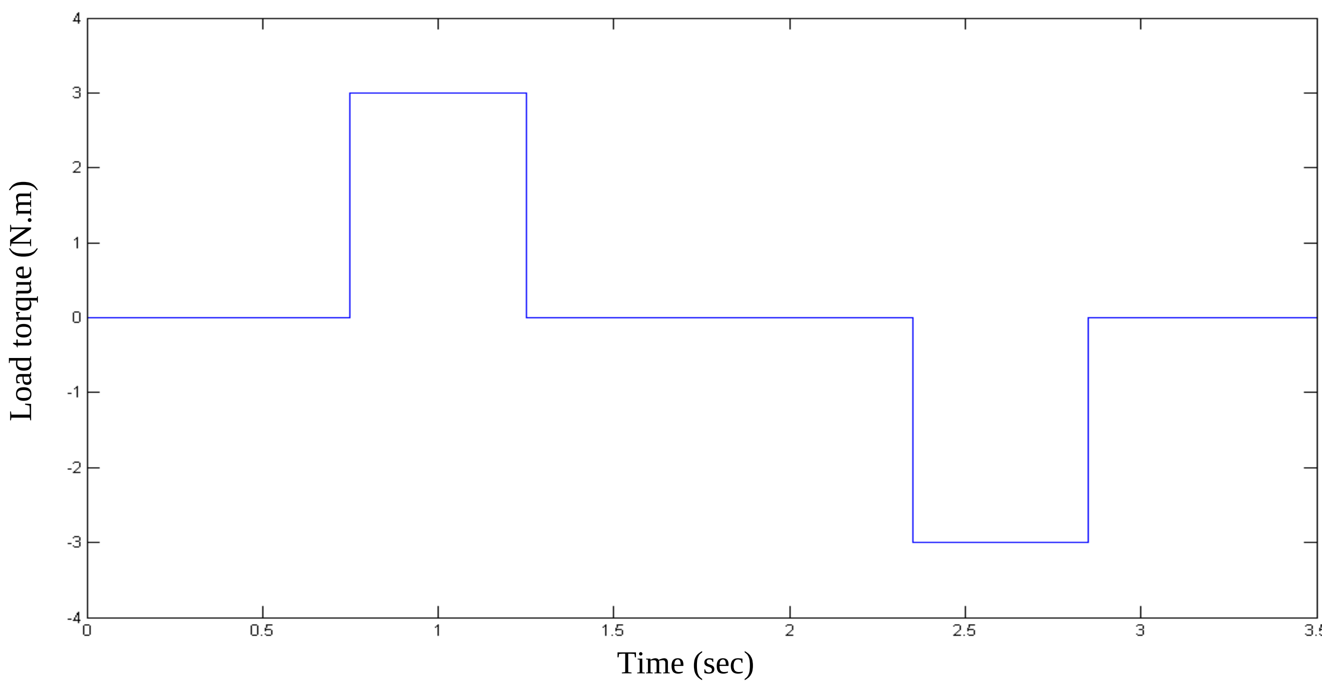


Fig. 4.2 Load torque trajectory

For sliding mode controller -

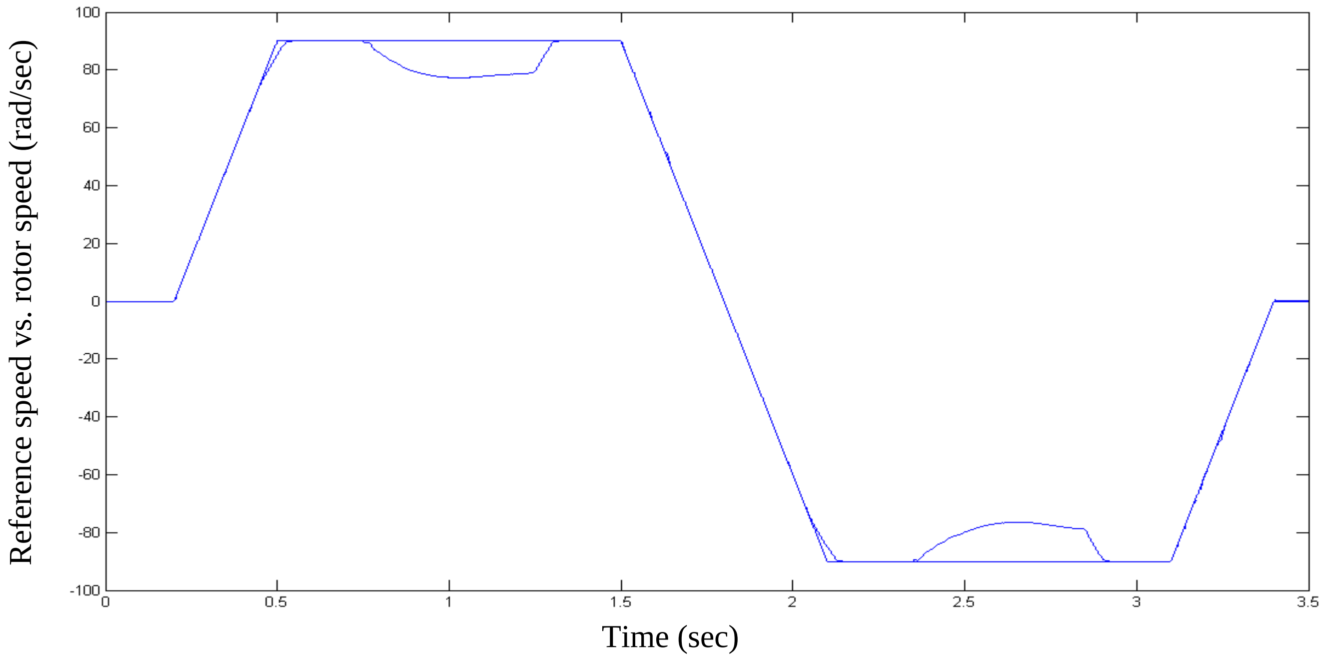


Fig. 4.3 Reference speed vs. rotor speed curve

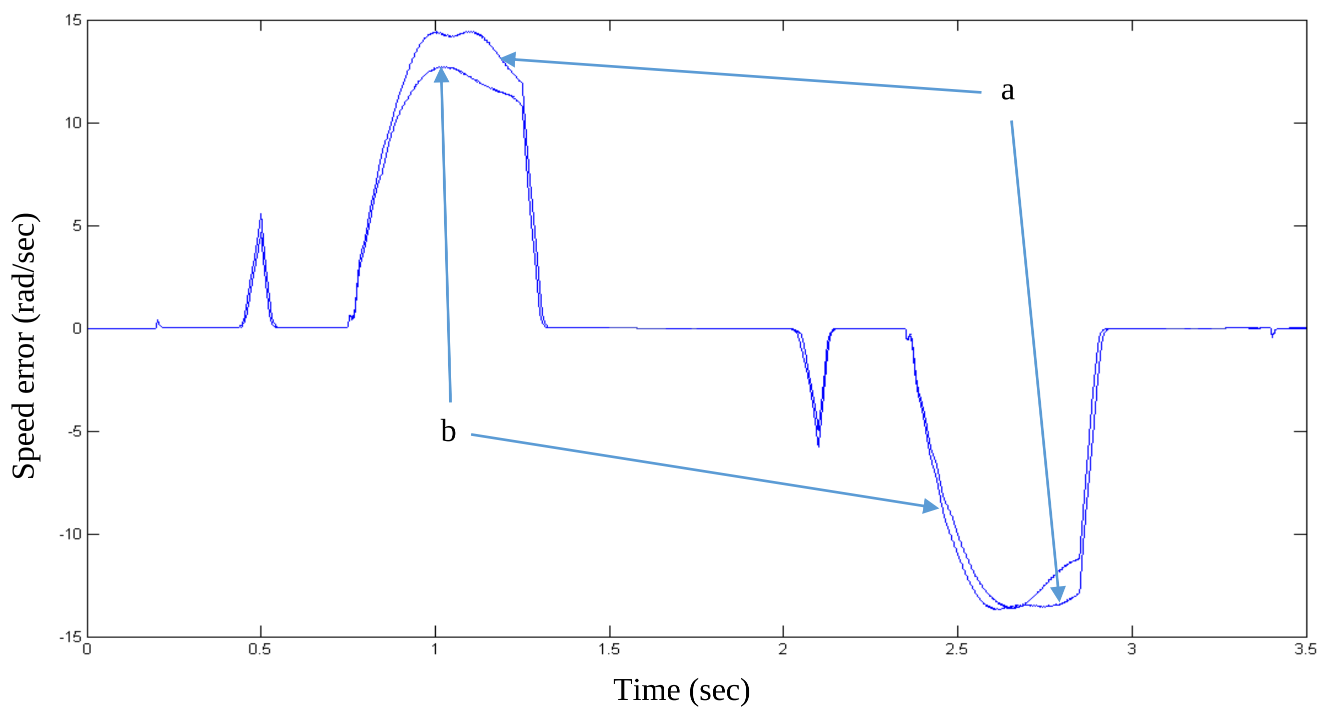


Fig. 4.4 Speed error for sliding mode controller with (a) standard inertia and (b) high inertia load

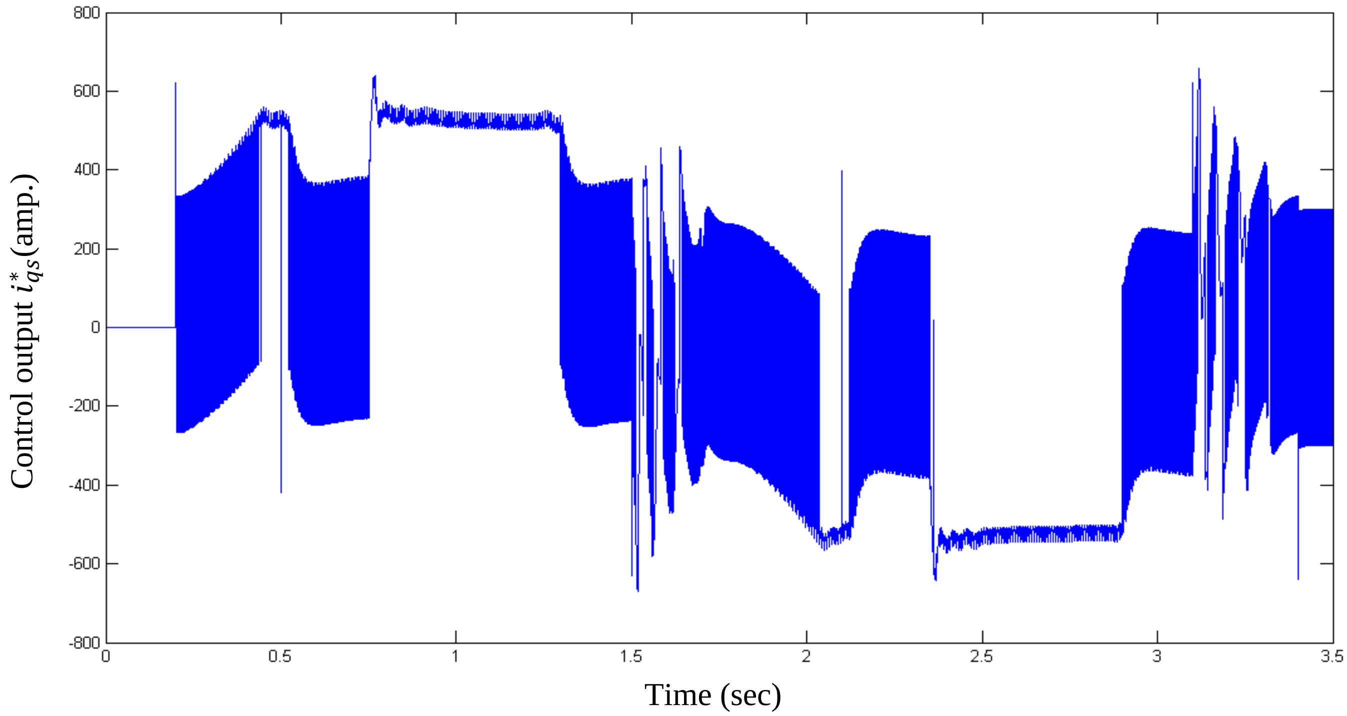


Fig. 4.5 Control output for sliding mode controller with standard inertia load

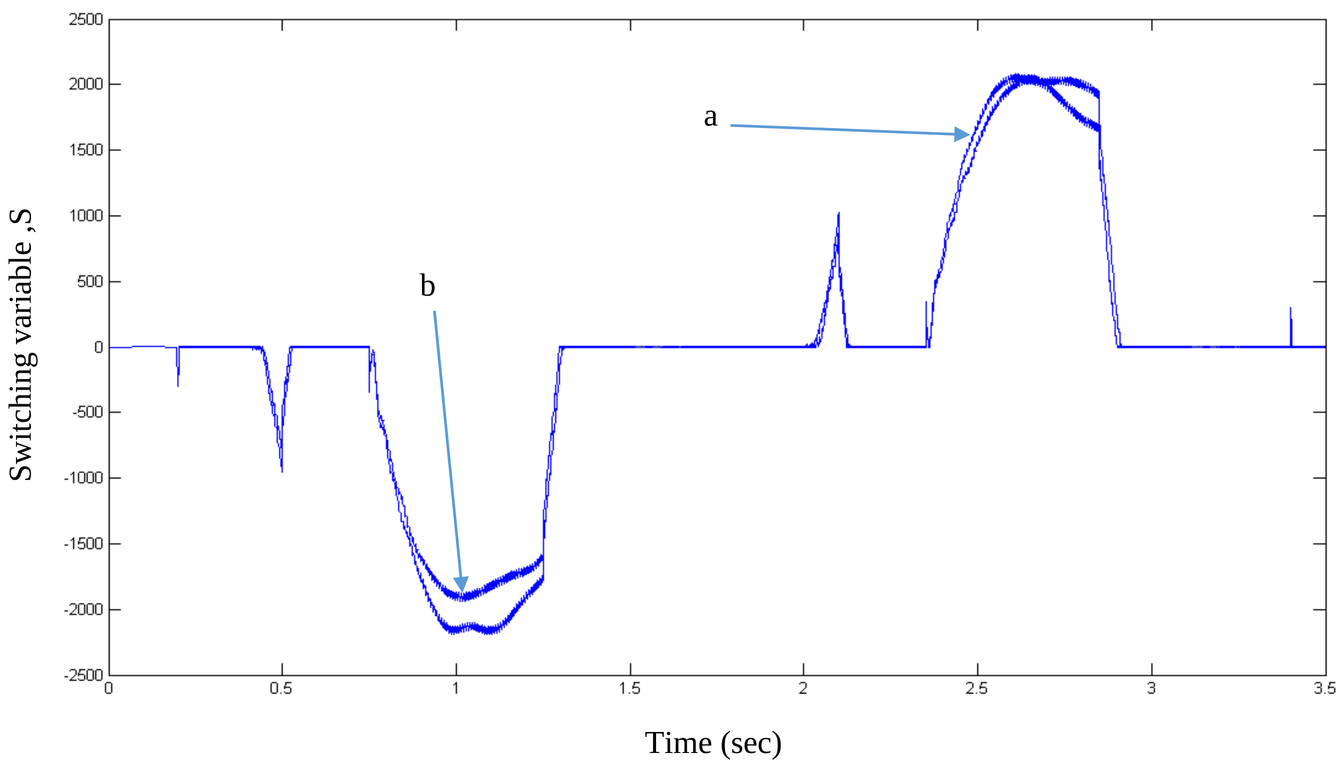


Fig. 4.6 Switching variable with (a) standard inertia and (b) high inertia load

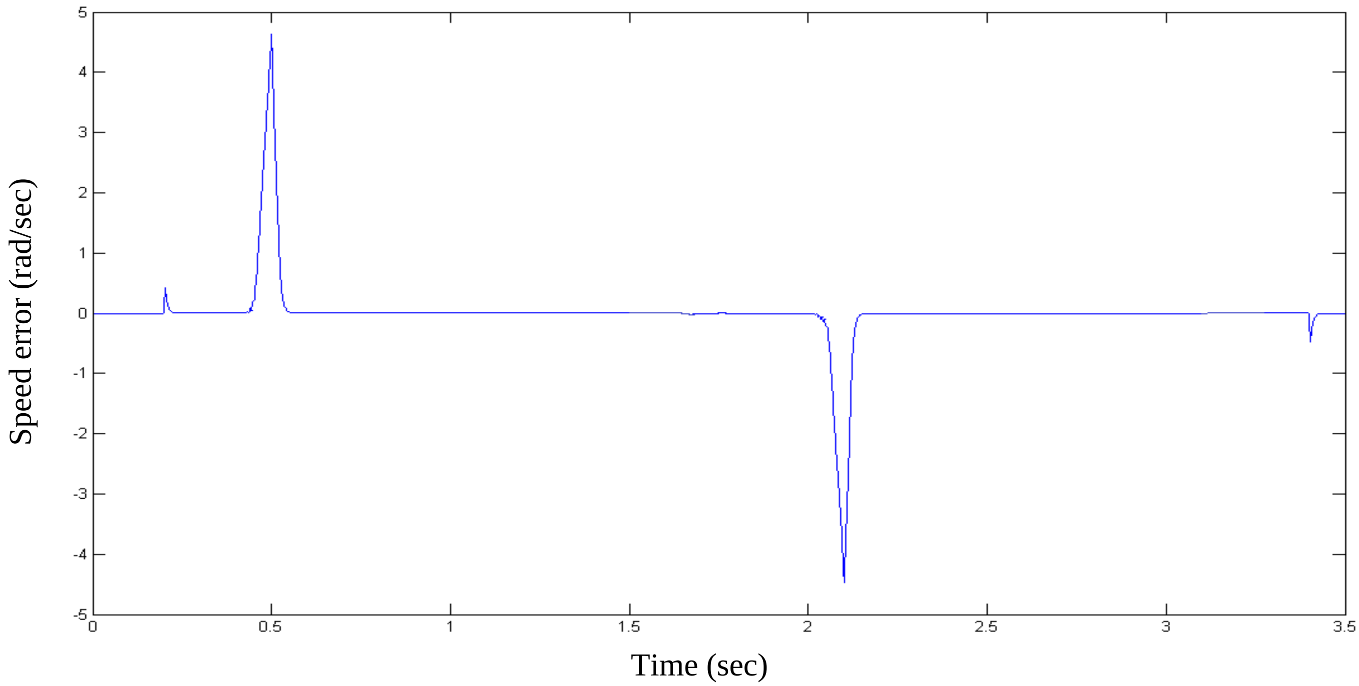


Fig. 4.7 No load speed error trajectory

For PI controller –

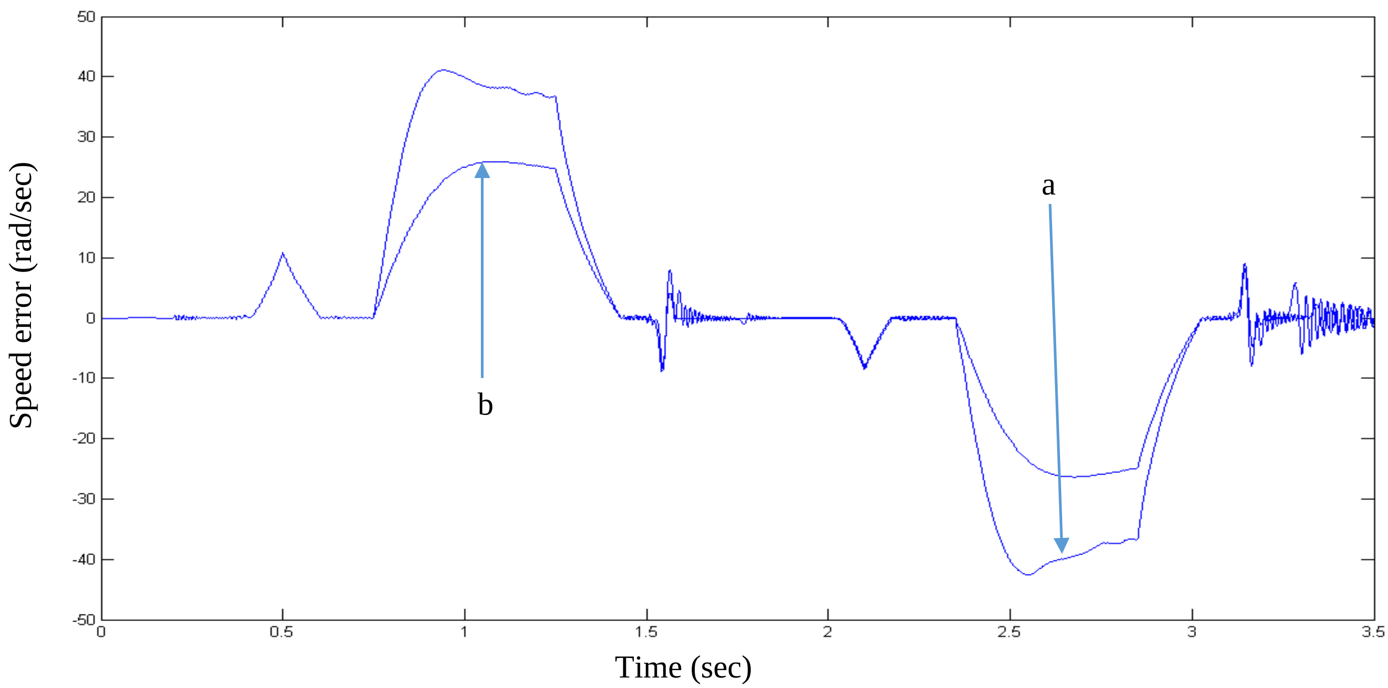


Fig. 4.8 Speed error for PI controller with (a) standard inertia and (b) high inertia load

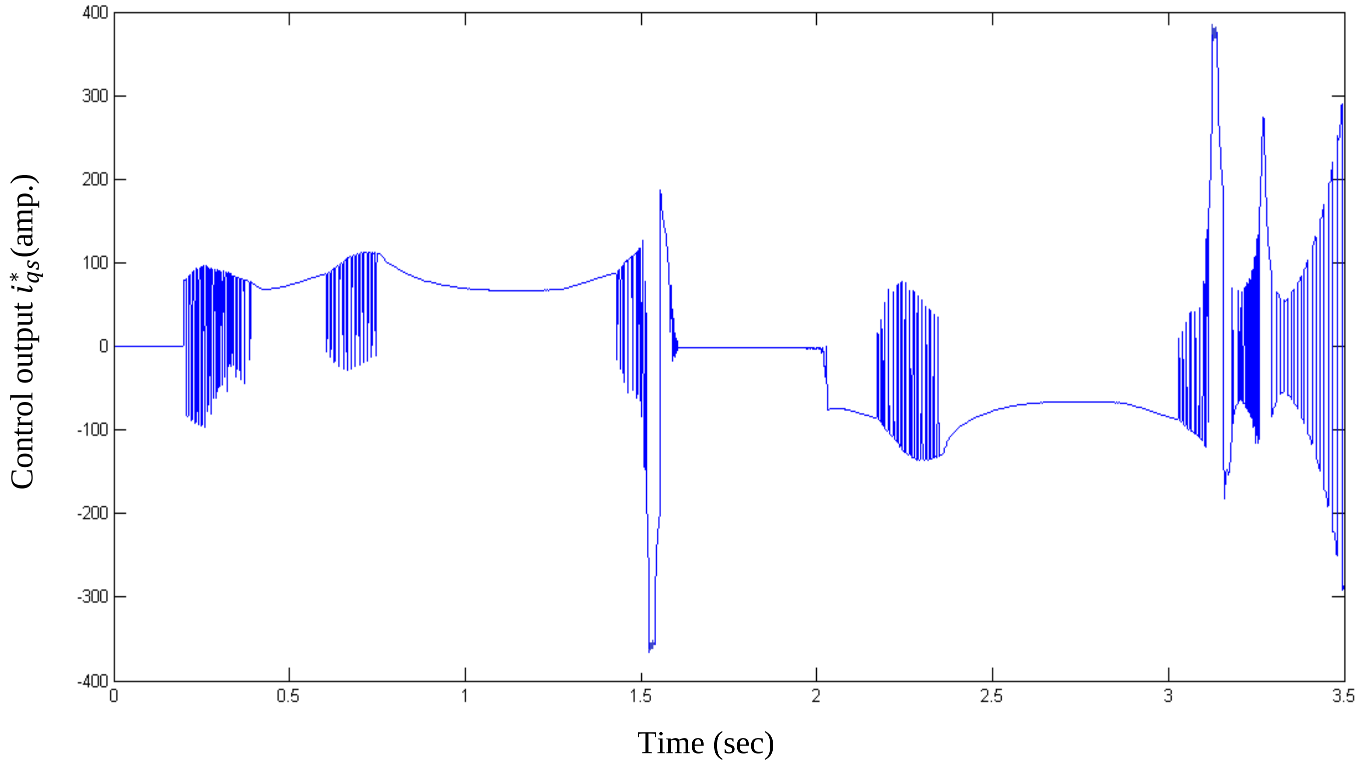


Fig. 4.9 Control output for PI controller with standard inertia load

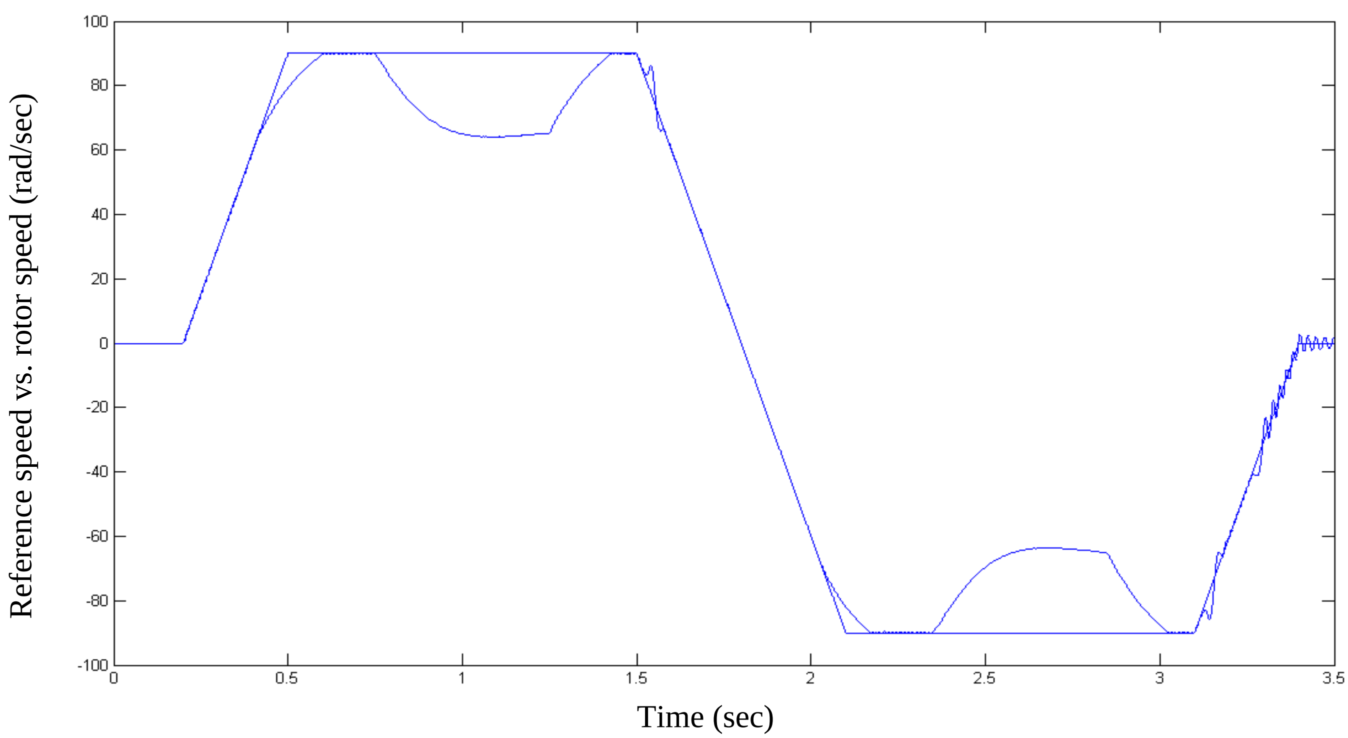


Fig. 4.10 Reference speed vs. rotor speed curve

4.3 Regulator performance

The drive system with sliding mode controller is subjected to step increments in speed command from 60 rad/s to 90 rad/s is demonstrated in Fig. 4.13. The speed error reaction is demonstrated in Fig. 4.14. The speed error response spike is beneath 30 rad/s. This shows sliding mode controller is additionally relevant for controller application.

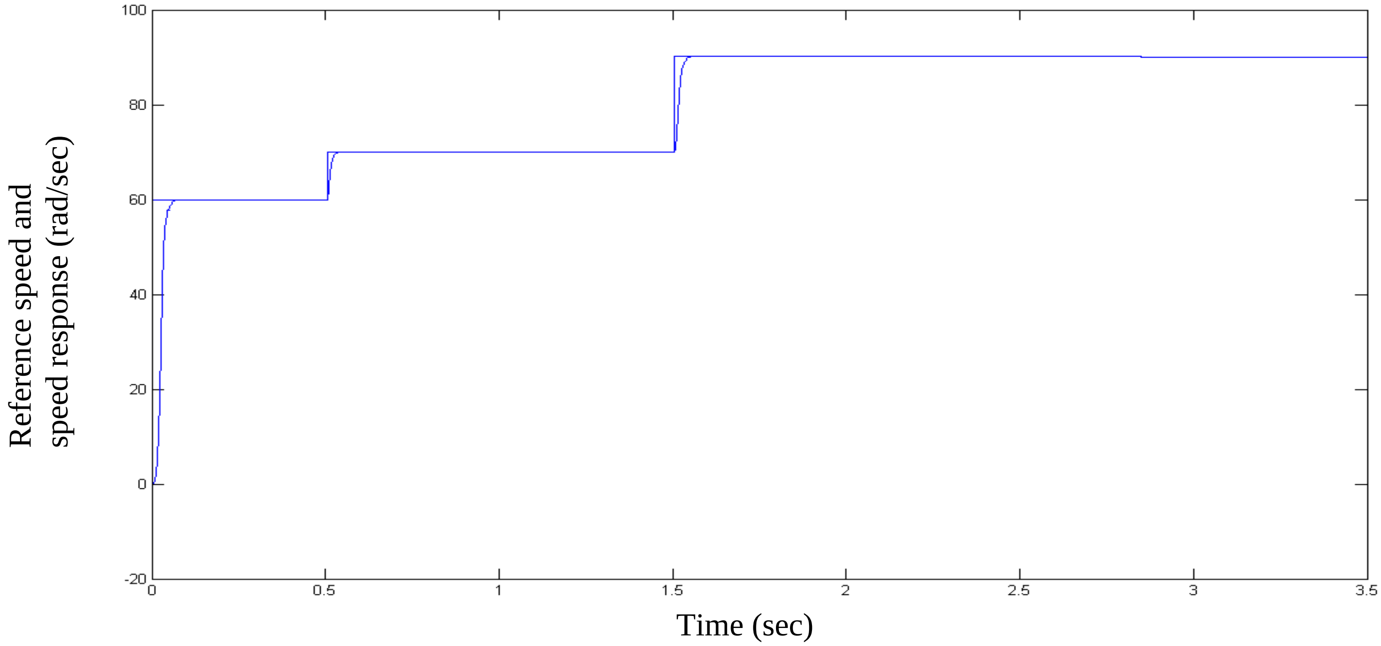


Fig. 4.11 Reference speed and speed response trajectory

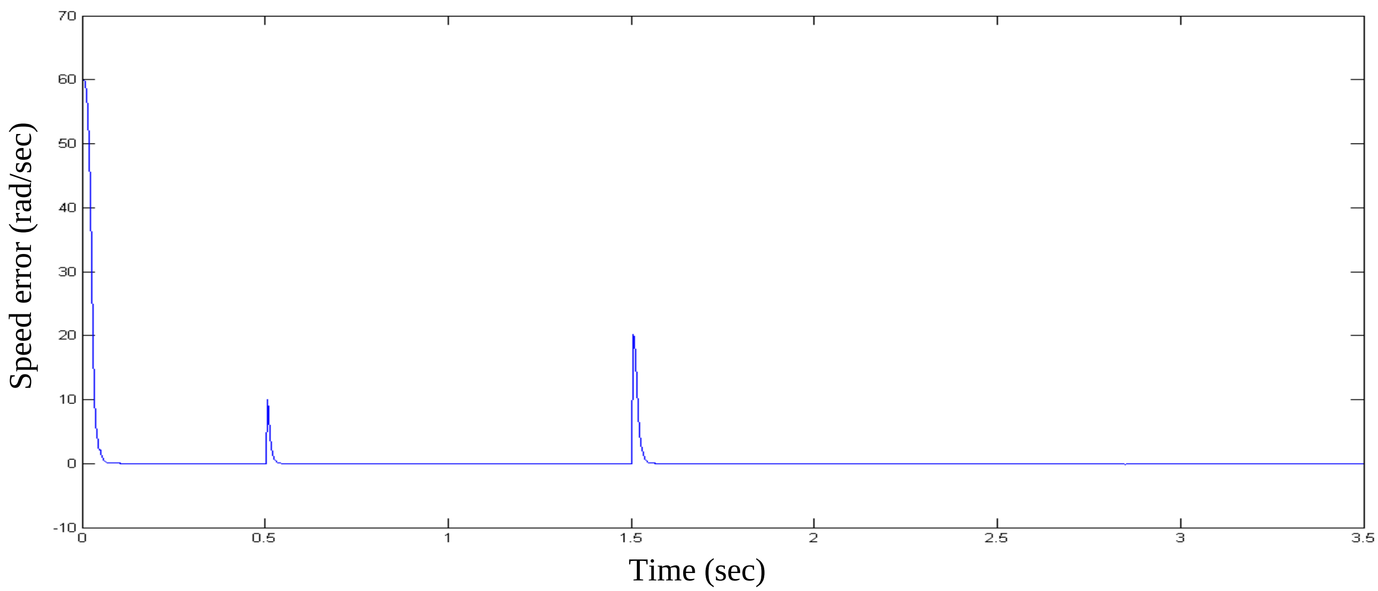


Fig. 4.12 Step changes in reference speed and speed response

4.4 Sensorless control

Sensorless control of the drive is implemented with speed estimation algorithm. The estimated speed is used in sliding mode controller. Fig.4.13 shown the estimated speed using rotor flux observer. Fig.4.14 shown the speed error between estimated speed and actual rotor speed. Fig. 4.15 shown the d – axis rotor flux trajectory and Fig. 4.16 shown the q – axis rotor flux trajectory which is synthesized from motor state equation.

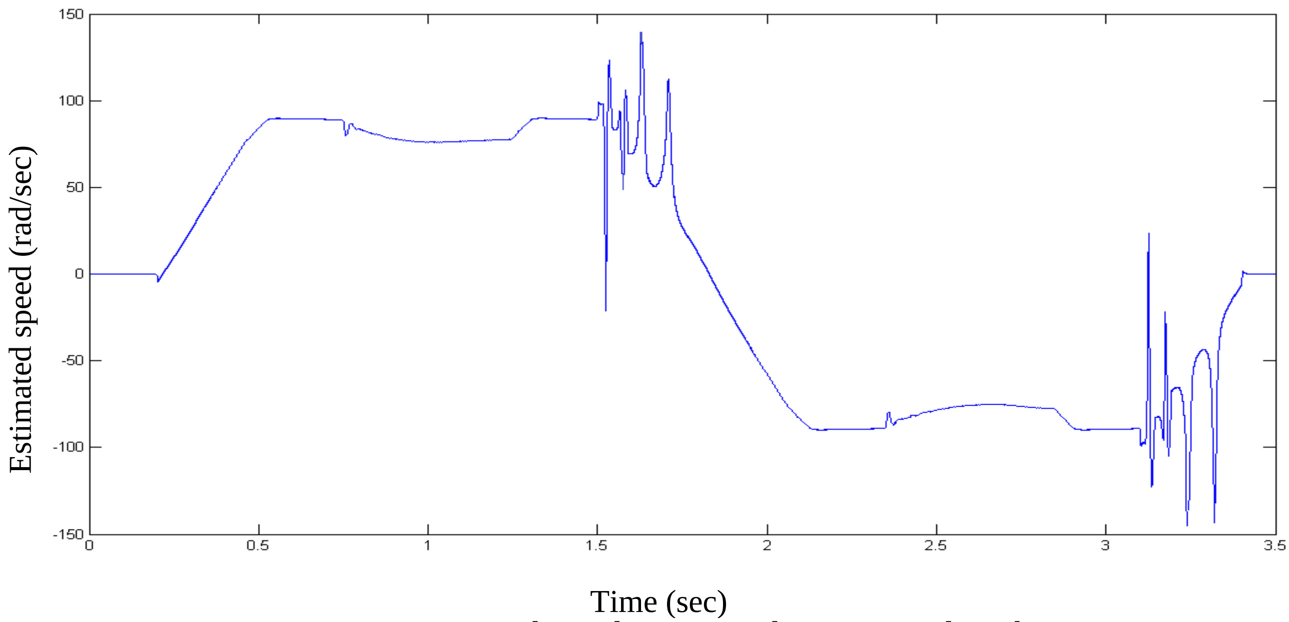


Fig. 4.13 Estimated speed using speed estimation algorithm

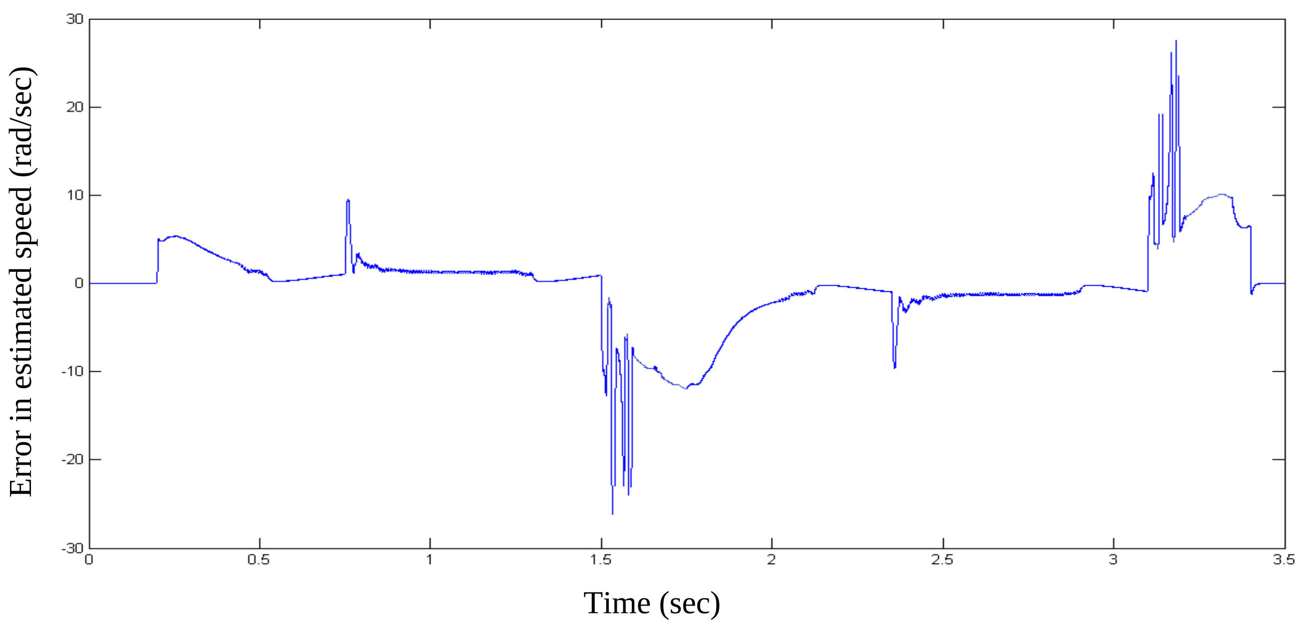


Fig. 4.14 Error in speed estimation

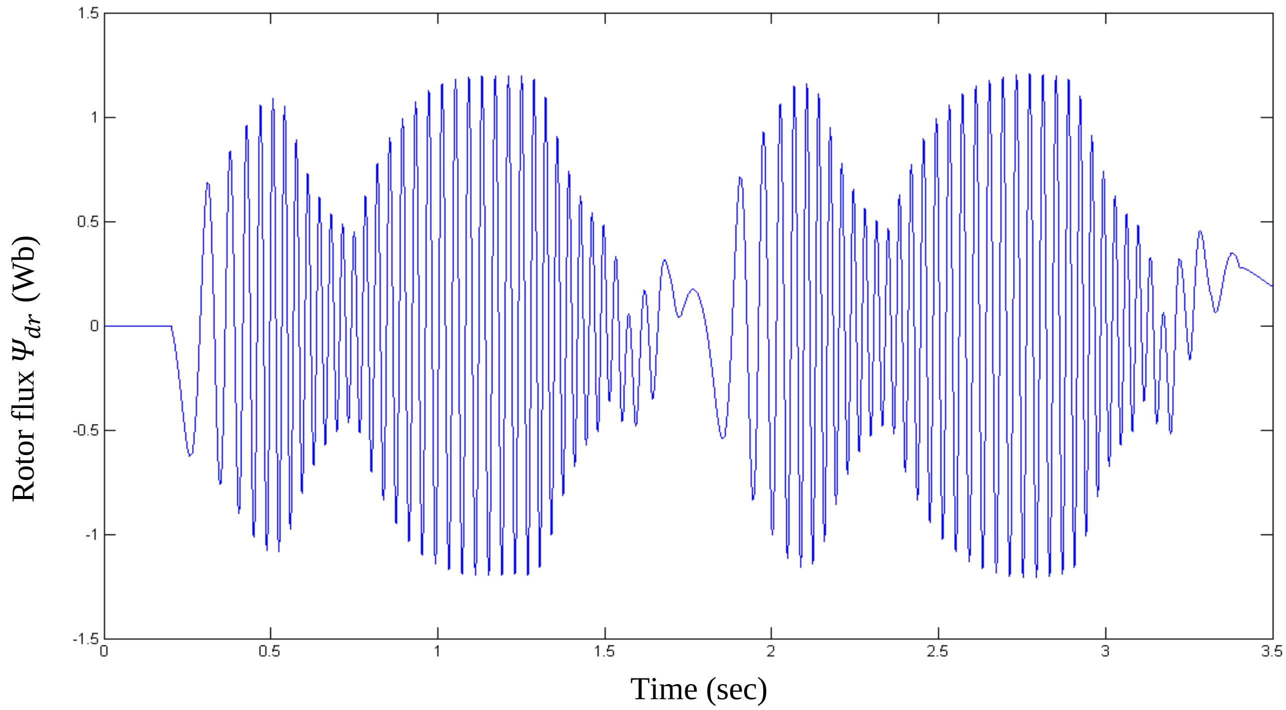


Fig. 4.15 Rotor d – axis flux trajectory

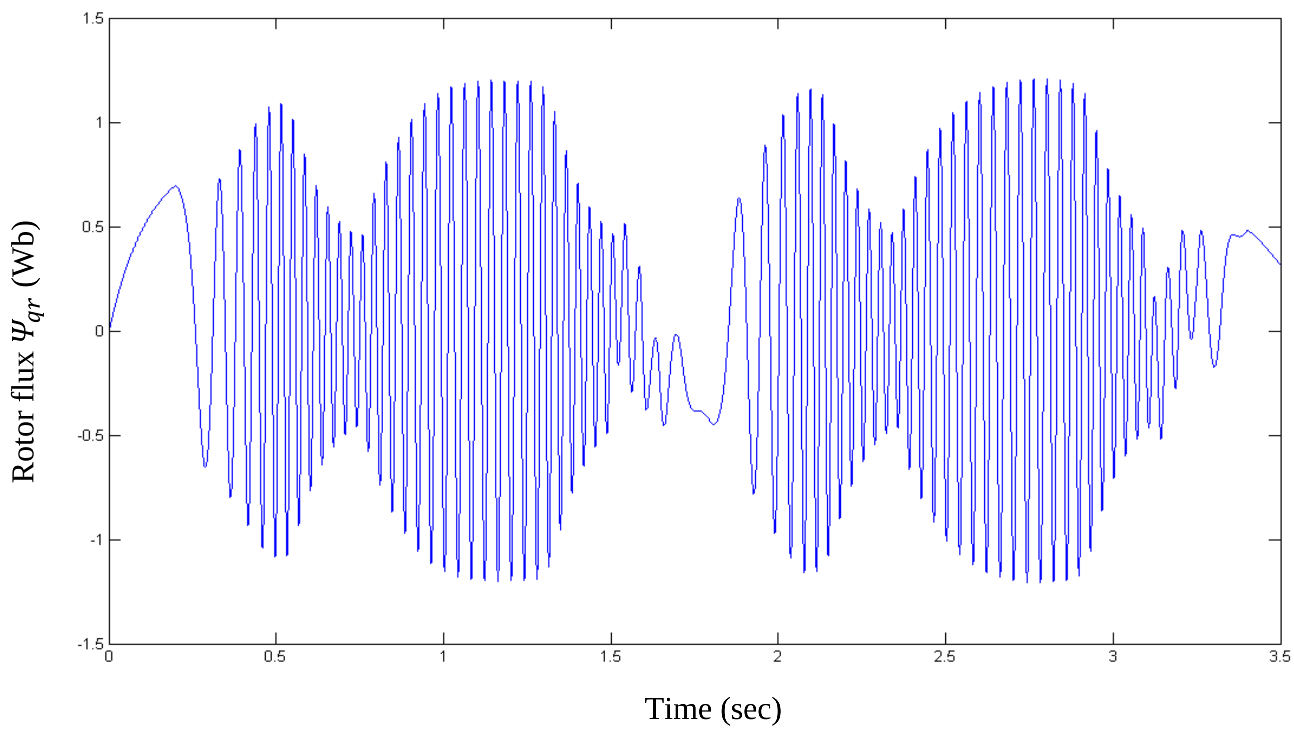


Fig. 4.16 Rotor q – axis flux trajectory

4.5 Conclusion

The simulation of sliding mode controller and PI controller using MATLAB/Simulink software is presented and the simulation results are shown that the error produced by fixed gain PI controller is larger than sliding mode controller. Then compare the simulation result of both controller for checking robustness and parameter variation of motor. Then discussed trajectory tracking performance and regulator performance of sliding mode controller and sensorless speed estimation is also simulated. The estimated speed contains ripples, due to mismatch of current and speed sampling times. But the speed response of drive system, with estimated speed being used in the control scheme, is acceptable.

CHAPTER-5

CONCLUSION AND FUTUREWORK

5.1 Conclusion

In this thesis, a sensorless sliding mode vector control has been presented. The rotor flux is estimated by means of a rotor flux observer based on the voltage and current equations of induction motor in stationary reference frame. The gain and bandwidth of sliding mode controller are designed. This work is based up on indirect vector control of induction motor and the speed control is analyzed by sliding mode controller as well as PI controller. Simulation results shows that the error produced by fixed gain PI controller is more than sliding mode controller. Robustness of sliding mode controller is also demonstrated. The regulator performance with sliding mode controller is better. Speed is estimated by rotor flux observer to achieve sensorless sliding mode control.

Then by simulation examples, it has been shown that the proposed control scheme perform reasonably well in practice and that the speed tracking objective is achieved under uncertainties in the parameters and load torque. The estimated speed contains ripples, due to mismatch of current and speed sampling times. But the speed response of drive system, with estimated speed being used in the control scheme, is acceptable.

5.2 Future work

Since the machine rating is small, the resistance variation effect is also small, so it has negligible effect. As a future work 2^{nd} order sliding mode controller can be implemented to reducing the chattering effect and parameter variation effect. Fuzzy logic principle can be applied to this controller to make it more robust and reliable.

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APPENDIX

RATING AND PARAMETERS OF THE INDUCTION MOTOR

Three phase, 50 Hz, 0.75 kW, 220 V, 3A, 1440 rpm

Stator and rotor resistances: $R_s = 6.37\ \Omega, R_r = 4.3\ \Omega$

Stator and rotor self inductances: $L_s = L_r = 0.26$ H

Mutual inductance between stator and rotor: $L_m = 0.24$ H

Moment of inertia of motor and load: J = 0.0088 kg.m^2

Viscous friction coefficient: B = 0.003 N.m.s/rad

For speed loop PI controller –

$K_p1 = 0.35$ $\quad\quad$ $K_I1 = 1$

$K_p2 = 500$ $\quad\quad$ $K_I2 = 24500$